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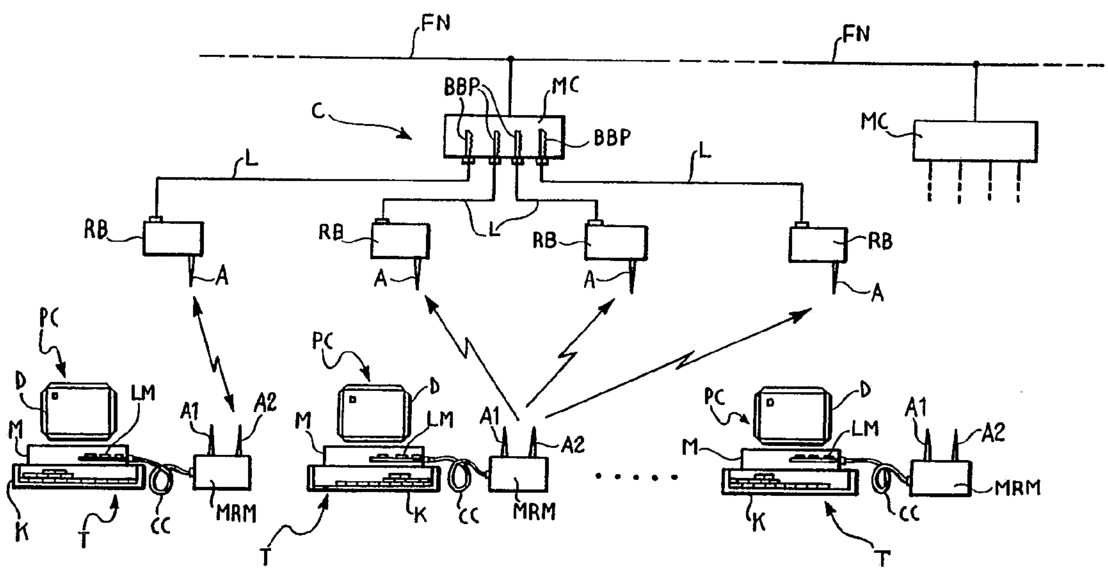

(57) Abstract

The network enables data to be communicated by radio, in accordance with the DECT standard, between the data terminals (PC) of a plurality of user stations (T), by means of a fixed central control device (C). Each user station (T) is associated with a mobile radio transmitter/receiver module (MRM) which is separate and distinct from the data-terminal (PC), and with an adaptor device (LM) which acts as an interface between the data terminal (PC) and the radio module (MRM) and which is physically incorporated in the data terminal (PC) and is connected to the radio module (MRM) by a flexible multicore cable (CC). The central control device (C) includes a multiplicity of fixed radio modules or bases (RB) and a fixed concentrator (MC) which is connected to the fixed radio bases (RB) by connecting lines (L).

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Cordless local area network (radio LAN) with central unit

The present invention relates to a local area network (a LAN) and, more specifically, to a network of the cordless (or wireless) type, particularly a network which operates in accordance with the DECT standard to enable data to be communicated by radio between a plurality of user stations each comprising a respective data terminal, by means of a fixed central control device.

Local networks have become increasingly widespread in the informatics and telematics world for short-range connections for enabling the transmission and distribution of data and services between a plurality of users within the same area, for example, in the same building. A local network enables many data terminals of different kinds, such as personal computers (PCs), minicomputers, printers, etc. to be connected in an extremely flexible manner, enabling very fast transmission speeds of the order of hundreds of thousands of Kbits/sec.

Up to now, most local networks have been of the wired type, that is, of the type in which the connections between the user stations and the central control devices are formed entirely by wires.

The appearance on the market of portable computers such as laptop personal computers, has created a need for cordless LANs.

A cordless local area network reduces installation costs because it eliminates the need to install connecting cables. This type of network can also be

formed in situations in which it would be difficult or impossible to install connecting wires, such as, in buildings which do not have sockets for LANs or in which there are architectural constraints.

A cordless LAN network may represent the ideal solution in an organisation in which the positions of the user stations or the number of stations connected in the network are subject to frequent changes or modifications.

A cordless LAN also represents the ideal solution for organisations which are subject to frequent changes of location. In this case, it would in fact be neither practical nor economical to transfer a wired LAN.

Finally, as stated above, a LAN network enables data to be communicated even by portable personal computers, without limiting the mobility of these new devices.

The network according to the invention operates in accordance with the DECT (Digital European Cordless Telecommunications) standard developed by ETSI - the European Telecommunications Standards Institute - which defines the specifications for radio connections between users and a network in a private environment.

The DECT system operates in the band between 1880 MHz and 1900 MHz and provides for radio transmission by means of a hybrid time and frequency multiplex system.

The characteristics of the DECT standard are described, for example, in "Digital European Cordless Telecommunications Services and Facilities", ETSI DR/RES 3003, June 1991 and in "Data Services in DECT",

A. Bud, Fifth International IEE Conference on Land Mobile Radio, Warwick, December 1989.

The cordless local area network according to the invention is characterised in that the data terminal of each user station is associated with:

- a mobile radio transmitter/receiver module which is separate and distinct from the data-terminal, and

- a microprocessor adaptor device for acting as an interface between the data terminal and the associated mobile radio module, the adaptor device being incorporated physically in the data terminal and connected to the mobile radio module by a flexible multicore cable,

and in that the central control device includes:

- a multiplicity of radio modules or bases for installation in respective predetermined fixed positions and for transmitting/receiving packets of data to/from the radio module of one or more user stations, and

- a microprocessor concentrator (hub) which is intended to be installed in a fixed position and to be connected to the fixed radio bases and which is programmed to control the communications between the user stations by means of the radio bases, according to predetermined procedures and protocols.

Typically, the data-terminals of the user stations may, for example, be personal computers and the microprocessor adaptor device is conveniently produced

in the form of a "half-size"-format card or daughter board incorporated in the PC and connected to the bus thereof. The electrical supply for the adaptor is thus conveniently derived from the data-terminal bus.

Moreover, the transmitter/receiver radio module to advantage takes its electrical supply from the associated adaptor board by means of conductors which extend through the flexible multicore cable connecting it to the board.

To advantage, each user station radio transmitter/receiver module has two omnidirectional antennae for achieving space "diversity" to improve the characteristics of the radio connection with the fixed radio modules or bases.

Conveniently, but not necessarily, the fixed central control device may be arranged for connection to a fixed network, for example, an Ethernet network or a Token Ring or RS232 network.

Further characteristics and advantages of the invention will become clear from the following detailed description of a cordless LAN network operating in accordance with the DECT standard, the description being given with reference to the appended drawings, provided purely by way of non-limiting example, in which:

Figure 1 is a block diagram of the LAN network,

Figure 2 is a circuit diagram showing the structure of an adaptor and a mobile radio module associated with each data-terminal of the LAN network shown in Figure

1,

Figure 3 is a time/frequency diagram relating to the manner in which radio transmission is effected according to a hybrid TDM/FDM system in the LAN network of Figure 1, and

Figure 4 shows an example of a frame for an asymmetric multi-bearer connection which can be formed in the LAN network of Figure 1.

With reference to Figure 1, a cordless local area network LAN formed in accordance with the specifications of the DECT standard includes a plurality of user stations T and a fixed central control device, generally indicated C.

Each user station T includes a respective data terminal which, in general, may be constituted by any device, such as a processor, a printer, etc., which can send and/or receive digital data by means of a communications network. In the embodiment shown by way of example in Figure 1, the data terminals of the user stations T are constituted by personal computers PC having standard network and applications software of the LAN Manager type. The personal computers may, for example, be Olivetti 1/D33 devices, each including a keyboard K, a display screen D and a processing module M.

Each data terminal PC is connected to a respective mobile radio transmitter/receiver (transceiver) module indicated MRM, of a type conforming to the DECT specifications for the physical layer.

The processing module M of each data-terminal PC incorporates a respective microprocessor adaptor device, indicated LM. The microprocessor adaptor is suitable for acting as an interface between the respective data terminal and the associated mobile radio module MRM. For this purpose, as shown schematically in Figure 2, the microprocessor adaptor LM is connected to the data bus DB of the processing module M of the data terminal. The adaptor LM is also connected to the mobile radio module MRM associated with the data terminal by means of a multicore flexible cable CC (Figure 1 and 2).

The central control device C includes a multiplicity of fixed radio modules or bases FRM installed in respective predetermined fixed positions for transmitting/receiving packets of data to/from the mobile radio module MRM of one or more user stations T.

The radio bases RB are connected, for example, by electrical wires L, to a microprocessor concentrator MC which is installed in a fixed position and is programmed to control the communications between the user stations T by predetermined procedures and protocols, in accordance with the DECT standard, by means of radio connections established between the mobile radio modules MRM and the radio bases RB.

Preferably, but not necessarily, the concentrator MC may be arranged for connection to a fixed network FN, for example an Ethernet network or a Token Ring or RS 232 network. Concentrators MC of other local networks LAN may possibly be connected to the fixed network.

The integrated system described with reference to

Figure 1 can perform the function of an MAC (medium access control) level multi-port bridge to enable the user stations T to transmit and receive packets of data which are packaged in accordance with the DECT standard format and are exchanged by radio, by means of the fixed portion C of the system. This portion acts as a very rapid packet-switching system and directs the packets received towards the destination user stations or towards the wired network FN.

The system described operates in accordance with the DECT standard. The DECT standard connection between the user stations T and the fixed portion C of the system replaces only the MAC level of the Ethernet system.

By virtue of the lines L, the radio bases RB can be installed up to distances of the order of 100 metres from the concentrator MC. By carrying out functions, such as connection handover, which are provided for in the DECT standard, almost complete continuity of service between the two or more radio bases RB used can be established.

The concentrator MC may be constituted, for example, by an Olivetti M300 personal computer with an Intel 386Sx processor operating at 16 MHz.

This concentrator incorporates baseband processors BBP connected in an orderly manner to respective associated radio bases RB.

Conveniently, the baseband processors BBP of the concentrator MC and the interface adaptors LM of the user station T may be in the form of half-size format

PC circuit boards and, in practice, may conveniently have the same structure at the hardware level and be differentiated only at the software level. The structure of an interface adaptor LM of a user station will be described in greater detail below with reference to Figure 2.

The concentrator MC as a whole is responsible for controlling the entire system and, in particular:

- the functioning of the high levels of the DECT protocols,

- the control of the various resources of the network,

- the switching of the packets of data, and possibly

- the interfacing between the cordless network LAN and the wired network FN.

The high levels of the DECT protocols provide for services such as fast handover, user authentication and the creation of virtual connections which enable physical connections to be established without massive exchanges of data.

Before the merits of the structure of the functions of the LM devices and of the band base processors BBP are discussed further, some characteristics relating to the mobile radio modules MRM and to the radio bases RB will be set out.

Structurally, the modules MRM and RB are almost identical. As already stated they are transceivers conforming to the DECT specifications for the Physical

Layer. In accordance with the DECT specifications, the radio modules operate in the band between 1880 MHz and 1900 MHz on ten channels spaced at 1.728 MHz intervals.

Typically, the modules can instantaneously transmit a power of about 250 mW with an envisaged activity cycle according to the DECT standard of between 4% and 96%.

The modules can transmit signals modulated according to filtered Gaussian FSK which is a non-coherent version of GMSK in which BT = 0.5 (BT is the product of the bandwidth B of the filter used and the duration T of the individual symbol).

Radio communications between the MRM modules and the radio bases RM take place according to a hybrid time and frequency multiplex system (TDM/FDM) with double simplex and duplex connections.

Transmission takes place within time cycles or frames having durations d of (for example) 10 ms, divided (for example) into 24 slots, of which, in accordance with the DECT specifications, a first half (12) normally serve for transmissions from the radio bases RB to the portable radio modules MRM and the second half (12) for transmissions in the opposite direction.

Figure 3 shows the grid of the slots (240) available with ten channels for each frame. In the grid, the time t is indicated on the abscissa and the frequency f is indicated on the ordinate. The frequencies associated with the ten channels are indicated f_1-f_{10} and the slots into which each individual frame is divided are numbered 1-24.

With frames each of 10 ms divided into 24 slots, each slot has a duration of 416.667 μs of which 364.667 μs can be used for a packet of data and 51 μs as a time interval (a guard space).

Conveniently, a time-division duplex (TDD) is used for duplex connections and slots at all the frequencies are used for multiple connections.

The radio modules MRM and RB therefore need to be able to retune themselves between two channels at opposite ends of the band and to switch between transmission and reception within the time interval (the guard space) between two slots.

The receiving portions of the radio modules MRM and of the radio bases RB conveniently have superheterodyne architecture with a single conversion stage.

As is clear from Figure 1, each radio base RB has a respective antenna A and the mobile radio modules MRM of the user stations each have two antennae A1 and A2 for achieving space diversity in order to improve the quality of the radio connections.

In the embodiment shown in Figure 2, each interface device LM associated with each data terminal includes a main microprocessor 50 and a signal processor 51.

The main microprocessor 50 which is constituted, for example, by a V40 device produced by Nippon Electric Company, can converse with the bus DB of the associated data terminal by means of a dual-port RAM memory 52 and with the other microprocessor 51 by means of another dual-port RAM memory 53.

The microprocessor 50 is associated with a program memory 54, for example, of the EPROM type and a RAM buffer memory 55 for the data.

The microprocessor 50 and the memory 55 are associated with a device 56 for controlling the interfacing with the memory and decoding the I/O ports. This device is conveniently formed as a large-scale integration ASIC integrated circuit (an application-specific integrated circuit).

The microprocessor 51 is a device for processing digital signals, for example, a TMS320 device produced by Texas Instruments and is programmed to control low-level MAC functions such as the formatting and deformatting of the frames and of the slots, the synchronisation of slots and frames, the detection of errors, the scanning of the communication channels, etc.

The processor 51 is also connected to a device 57 which extracts the clock signals from the signals received by the mobile radio module MRM and generates the timing signals and also effects any coding for protecting the data to be transmitted. The device 57 may also conveniently be produced in the form of a single ASIC integrated circuit.

This device is associated with a buffer 58 which acts as a protection latch. The processor 51 is connected by means of the buffer and the multicore cable CC to a device 59 within the mobile radio module MRM for controlling the radio transmission/reception circuits 60. The device 59 may also conveniently be produced in the form of an ASIC integrated circuit.

Conveniently, the device LM draws its electrical supply from the bus DB of the data terminal, for example, by means of the two conductors indicated 60 in Figure 2. Moreover, the electrical supply of the mobile radio module MRM to advantage is derived from that of the adaptor device LM, for example, by means of two conductors indicated 61 in Figure 2, which extend through the multicore interconnecting cable CC.

As stated above, from a hardware point of view, the baseband processors BBP of the concentrator device MC have the same structure as the logic modules LM fitted in the data terminals of the user stations T. In fact most of the functions of the baseband processors correspond to functions carried out by the modules LM. These functions include, in particular:

- the creation and dismantling of the slot structures,

- the creation and dismantling of logic channels,

- the monitoring of the free channels in the incoming communications,

- the propagation of "connectionless" and "paging" messages,

- handover between logic and "inter-cell" channels,

- the control of rapid procedures for detecting and correcting errors.

The interface adaptors LM of the data terminals are arranged also to perform the following functions:

- the creation and updating of a map of the usage of the physical communications channels and the selection of the channel for each connection to be established, and

- the decision to effect either intra-cell or inter-cell handover and the initiation thereof.

The adaptor modules LM also act as interfaces between the DECT environments and the applications environments of the respective data terminals. The module LM thus responds to the network operating system (the LAN manager) resident in the data terminal in exactly the same manner as an Ethernet network adaptor by means of a Microsoft Network Driver Interface Specification standard interface.

Two critical requirements for the application of the DECT specifications in a local area network LAN are the need to use the spectral resources with maximum efficiency and the need to minimise the delay introduced by the DECT. In order to achieve both these objectives, it is necessary to use specific protocols.

Since the data traffic is characterised by short transactions interposed between long silences it is inconceivable to keep the connections between the user stations and the radio bases open permanently since they would be massively underused. The radio connections are therefore established in the network only when there are data to transmit and are closed during periods of inactivity in order to free radio channels for use by other users.

For this purpose, the main processor 50 of each module LM is programmed to operate in the following manner.

Each time data are admitted to the buffer memory 55 for transmission by means of the associated mobile radio module MRM, the main microprocessor 50 sets up a radio connection by means of the microprocessor 51 (with a radio base determined in the manner which will be described below and with the use of slots of a channel or frequency determined in the manner which will also be described below). The radio connection thus opened is maintained throughout the time necessary for the transmission of the data in the memory 55. After the data have been transmitted the radio connection is not closed immediately but is kept open for a predetermined period of time. Conveniently, the main microprocessor 50 is arranged to process a short-term statistic relating to the communications traffic of the data terminal (for example, over a period of half an hour or an hour). The radio connection opened for the transmission of data is then closed with a delay after the moment at which the transmission of data is completed, the delay being determined adaptively on the basis of the mean traffic which has affected the data terminal. This reduces useless periods since, in many cases, it is not necessary to reopen the radio connection when a further flow of data arrives for transmission.

In order to select the radio base with which to establish the connection, each user station adaptor module LM operates in the following manner.

In accordance with the DECT standard, the main microprocessor 50 of the adaptor (LM) of each user

station is arranged cyclically to scan all the slots of all the channels by means of the associated mobile radio module MRM in order to detect the level of the signal emitted by each fixed radio base RB in each slot for each channel or frequency. On the basis of the levels of the signals thus detected, the microprocessor 50 can establish which is the nearest fixed radio base RB. The processor is also arranged, during the scanning, to decode the signals indicative, for each slot, of the radio base RB which is (possibly) active.

By virtue of this "mapping", in order to transmit data, the main processor 50 of the device LM of each user terminal can select the nearest radio base of which not all the slots are occupied at the time in question.

This procedure avoids futile attempts to establish a radio connection with a radio base which, although it is the nearest, is fully occupied at the time in question.

In accordance with the DECT standard, the baseband processors BBP of the concentrator device MC are arranged to scan the channels or frequencies f_1 - f_{10} cyclically by means of the associated radio bases RB. In particular, the scanning takes place in synchronism with the cyclical scanning effected by the devices LM of the user terminals. Moreover, the main processors 50 of the interface adaptor modules LM are arranged to carry out the scanning one channel in advance. In other words, if, in the course of their scanning, the fixed radio bases RB are "interrogating" the channel or frequency f_i, at the same moment, the mobile radio modules are "interrogating" the channel or frequency f_{i+1}.

This minimises the time needed to establish a radio connection between a user terminal and a fixed radio base.

Conveniently, the main processors 50 of the interface adaptors LM of the user stations and the baseband processors BBP of the concentrator MC are arranged to carry out the DECT Multibearer and Asymmetric Connection procedures in order to determine in which slot to transmit.

The multibearer procedure enables several slots (bearers) to be assigned simultaneously to the connection associated with a single user station. The bandwidth available for a user station may thus be increased from, for example, 32 kb/s duplex (single bearer) up to (theoretically), for example, 384 kb/s duplex with all twelve pairs of slots (12 bearers) in use.

Since the traffic in a local area network is typically very asymmetrical with the need to have considerable bandwidths available in one direction in particular, the DECT specifications include mechanisms which enable the uplink and downlink slots of a connection to be used in a single direction. A connection of this type must form part of a multibearer connection in which at least one other connection remains duplex to provide a route for control data in the opposite direction. The result is that a user can access almost the whole of the bandwidth (352 kb/s) by occupying half of the slots as shown in Figure 4, which relates to an asymmetric multibearer connection (5, 1).

Finally, the software used in the network LAN

conveniently includes procedures for detecting and correcting errors in accordance with the DECT specifications. The specifications provide for, at the level 2 (MAC/DLC), some mechanisms which have been developed appropriately for this purpose, and the main characteristics of which are the following:

- the MAC provides a service defined as an "Ip" (a protected information channel) with a throughput of 25.6 kb/s per connection and an error factor of 10^{-5}; this service is based on a retransmission mechanism which is quick and simple by virtue of the use of a single window packet;

- the DLC (data link control) provides a service defined as "Frame Relay" which protects the data against any errors introduced during handover and connection changes and against residual errors of the Ip channel.

Naturally, the principle of the invention remaining the same, the forms of embodiment and details of construction may be varied widely with respect to those described and illustrated purely by way of non-limiting example, without thereby departing from the scope of the present invention.

CLAIMS

1. A cordless local area network (LAN) for enabling data to be communicated by radio between a plurality of user stations (T) each comprising a respective data terminal (PC), by means of a fixed central control device (C), in accordance with the DECT standard,

characterised in that the data terminal (PC) of each user station (T) is associated with:

- a mobile radio transmitter/receiver module (MRM) which is separate and distinct from the data terminal (PC), and

- a microprocessor adaptor device (LM) for acting as an interface between the data terminal (PC) and the associated mobile radio module (MRM), the adaptor being incorporated physically in the data terminal (PC) and connected to the mobile radio module (MRM) by a flexible multicore cable (CC),

and in that the central control device (C) includes:

- a multiplicity of radio modules or bases (RB) for installation in respective predetermined fixed positions and for transmitting/receiving packets of data to/from the mobile radio module (MRM) of one or more user stations (T), and

- a microprocessor concentrator (MC) which is intended to be installed in a fixed position and to be connected, by connecting lines (L), to the fixed radio bases (RB) and which is programmed to control the

communications between the user stations (T) by means of the radio bases (RB), according to predetermined procedures and protocols.

2. A local area network according to Claim 1, in which each data terminal (PC) includes a data bus (DB), the network being characterised in that the microprocessor adaptor (LM) associated with each data terminal (PC) includes:

- means (51-58) for activating/de-activating the radio connection,

- a buffer memory (55), and

- a main microprocessor (50) which is connected to the data bus (DB) of the data terminal (PC), to the buffer memory (55), and to the means (51-58) for activating/de-activating the radio connection, the main microprocessor (50) being arranged:

to control the exchange of data with the data terminal (PC) in a predetermined manner,

to admit to the buffer memory (55) the data which are to be transmitted by the associated mobile radio module (MRM), and

to pilot the activating/de-activating means (51-58) in a manner such as to activate a radio connection each time data are stored in the memory (55) and to keep the radio connection open for a predetermined period of time after the transmission of the data in the memory (55) has been completed.

3. A local area network according to Claim 2, characterised in that the main microprocessor (50) is arranged to pilot the activating/de-activating means (51, 57, 58) in a manner such that, upon completion of the transmission of the data in the memory (55), the radio link is kept open for a period of time which is determined adaptively on the basis of a communications traffic statistic relating to the data terminal (PC) and calculated over a predetermined period of time.

4. A local area network according to any one of the preceding claims in which, in accordance with the DECT standard, the radio communications between the mobile radio modules (MRM) and the fixed radio bases (RB) take place according to a mixed time and frequency multiplex system (TDM, FDM) on n channels or frequencies (f_1 - f_{10}) within a predetermined band with time cycles (frames) of predetermined duration, divided into a predetermined number (2m) of time slots, and in which the main microprocessor (50) of the adaptor (LM) of each user station (T) is arranged:

- to scan all the 2m x n slots of all the n channels (f_1 - f_{10}) cyclically by means of the associated mobile radio module (MRM) and to detect the level of the signal emitted by each fixed radio base (RB) in each slot for each channel or frequency and thus to determine which radio base (RB) is nearest the user station (T),

the network being characterised in that the main microprocessor (50) is also arranged, during the scanning, to decode the signals indicative of the radio base (RB) which is (possibly) active in each slot and to select - in order to transmit data - the nearest

radio base (RB) for which not all the slots are occupied.

5. A local area network according to any one of the preceding claims, characterised in that the concentrator device (MC) includes a multiplicity of baseband processors (BBP) each of which is associated with and connected to a respective fixed radio base (RB) and is arranged to perform the functions up to level 2 of the hierarchy of DECT protocols.

6. A local area network according to Claim 5, characterised in that the baseband processors (BBP) are arranged to scan the transmission channels or frequencies (f_1 - f_{10}) cyclically, in accordance with a predetermined sequence, by means of the associated radio bases (RB), and in that the main processors (50) of the adaptors (LM) of the user stations (T) are arranged to scan the transmission channels or frequencies (f_1 - f_{10}) in synchronism with the baseband processors (BBP) but one channel in advance thereof.

7. A local area network according to any one of Claims 2 to 6, characterised in that the main processors (50) of the adaptors (LM) of the user stations (T) and the baseband processors (BBP) of the concentrator (MC) are arranged to effect the DECT multibearer and asymmetric connection procedures in order to determine the slots in which to transmit.

8. A local area network according to any one of the preceding claims, characterised in that the adaptor (LM) of each user station (T) is formed on a half-size format PC-AT circuit board.

9. A local area network according to one of Claims 5 to 8, characterised in that the baseband processors (BBP) are incorporated in the concentrator device (MC) and are supplied thereby.

10. A local area network according to any one of the preceding claims, characterised in that each mobile radio module (MRM) receives its electrical supply from the associated adaptor (LM) by means of the multicore cable (CC) which interconnects them.

11. A local area network according to any one of the preceding claims, characterised in that each mobile radio module (MRM) has a pair of antennae (A1, A2) for achieving space "diversity".

12. A local area network according to Claim 11, characterised in that each fixed radio base (RB) has a single antenna (A).

13. A local area network according to any one of the preceding claims, characterised in that the concentrator (MC) can be connected to a fixed network (FN) such as an Ethernet or Token Ring network and can converse therewith.

14. A local area network according to Claims 1 and 5, characterised in that the adaptors (LM) of the user stations (T) and the baseband processors (BBP) of the concentrator (MC) are formed by circuit boards which are identical from the hardware point of view but which are differentiated at the software level.

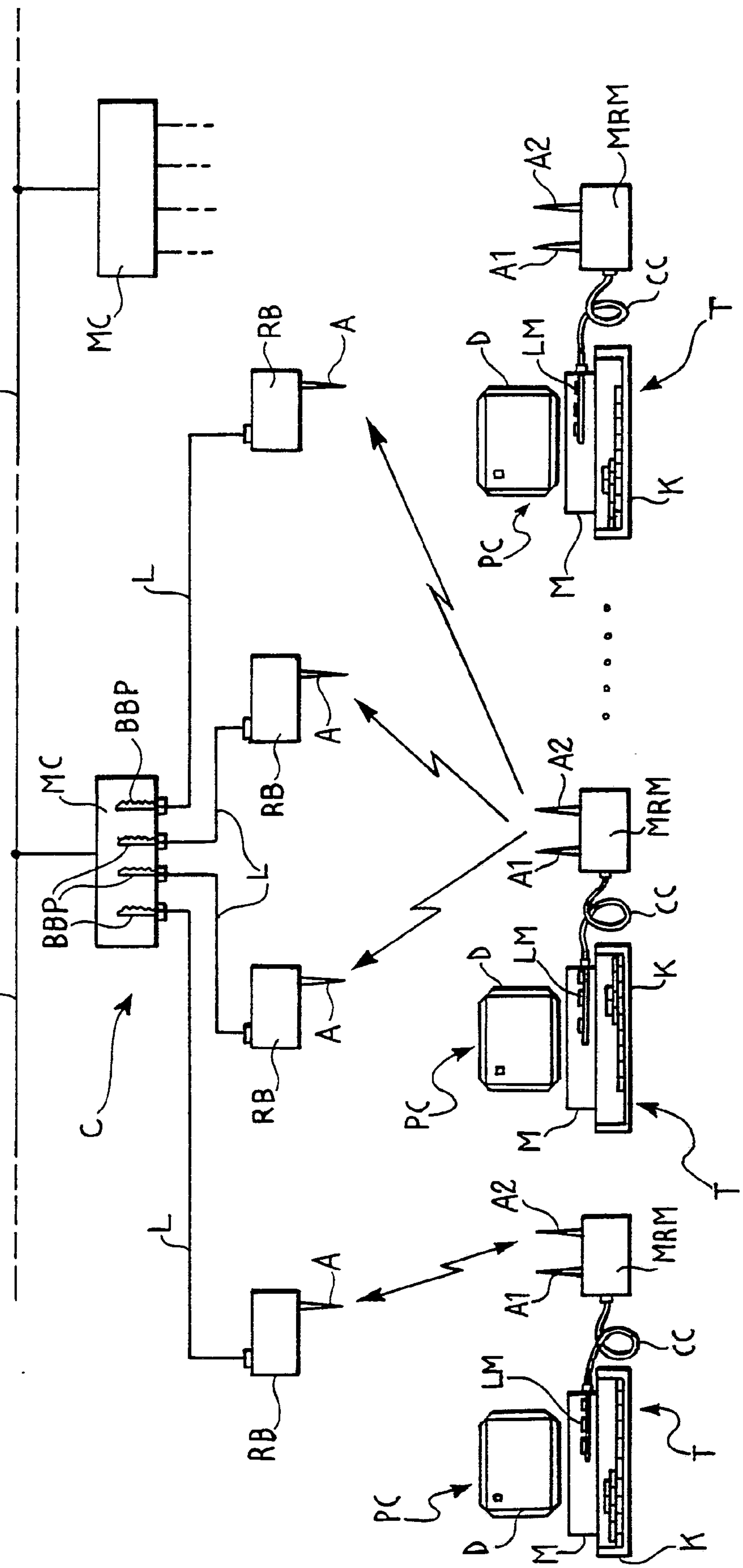
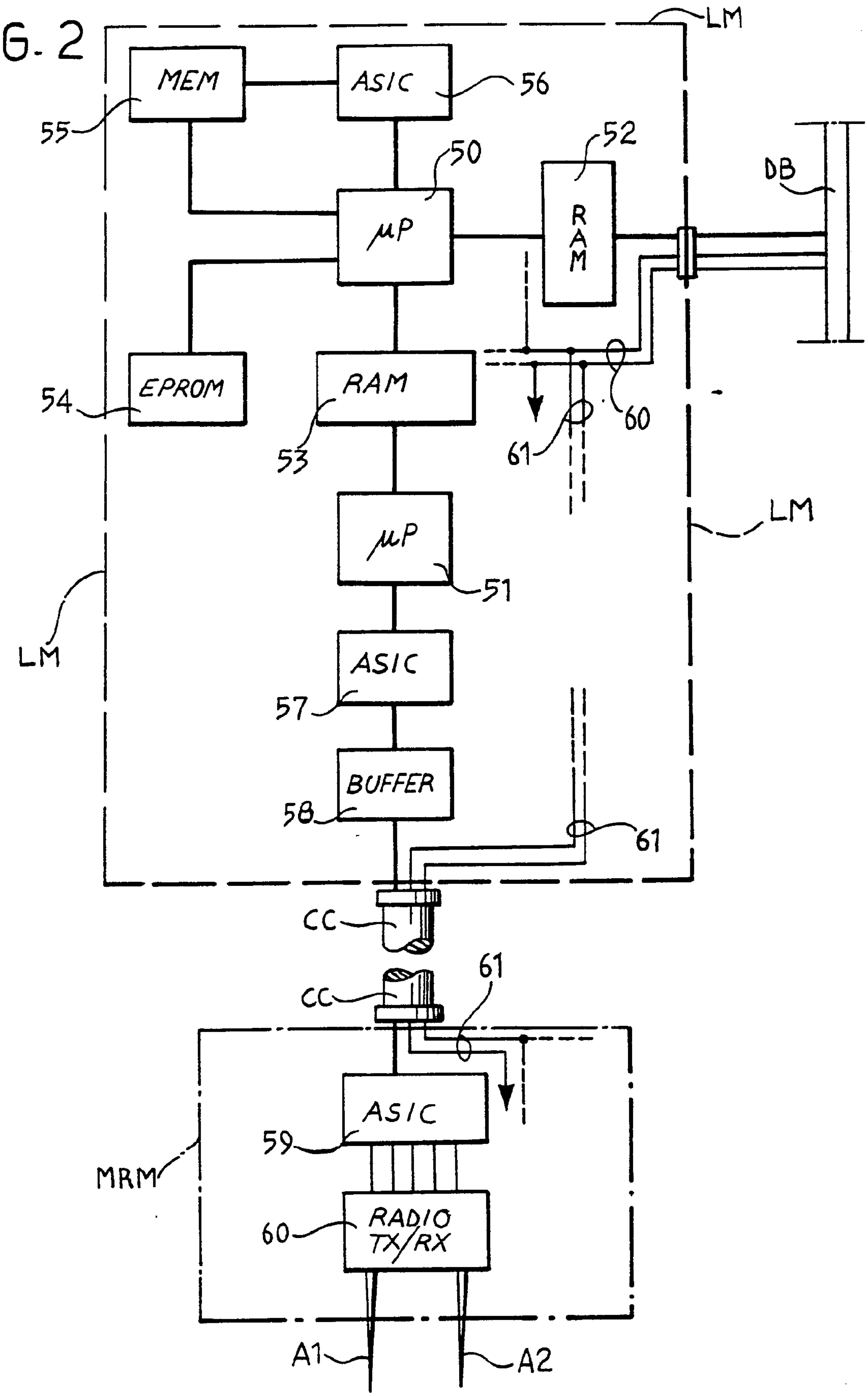


FIG. 1

FIG. 2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CC
CC
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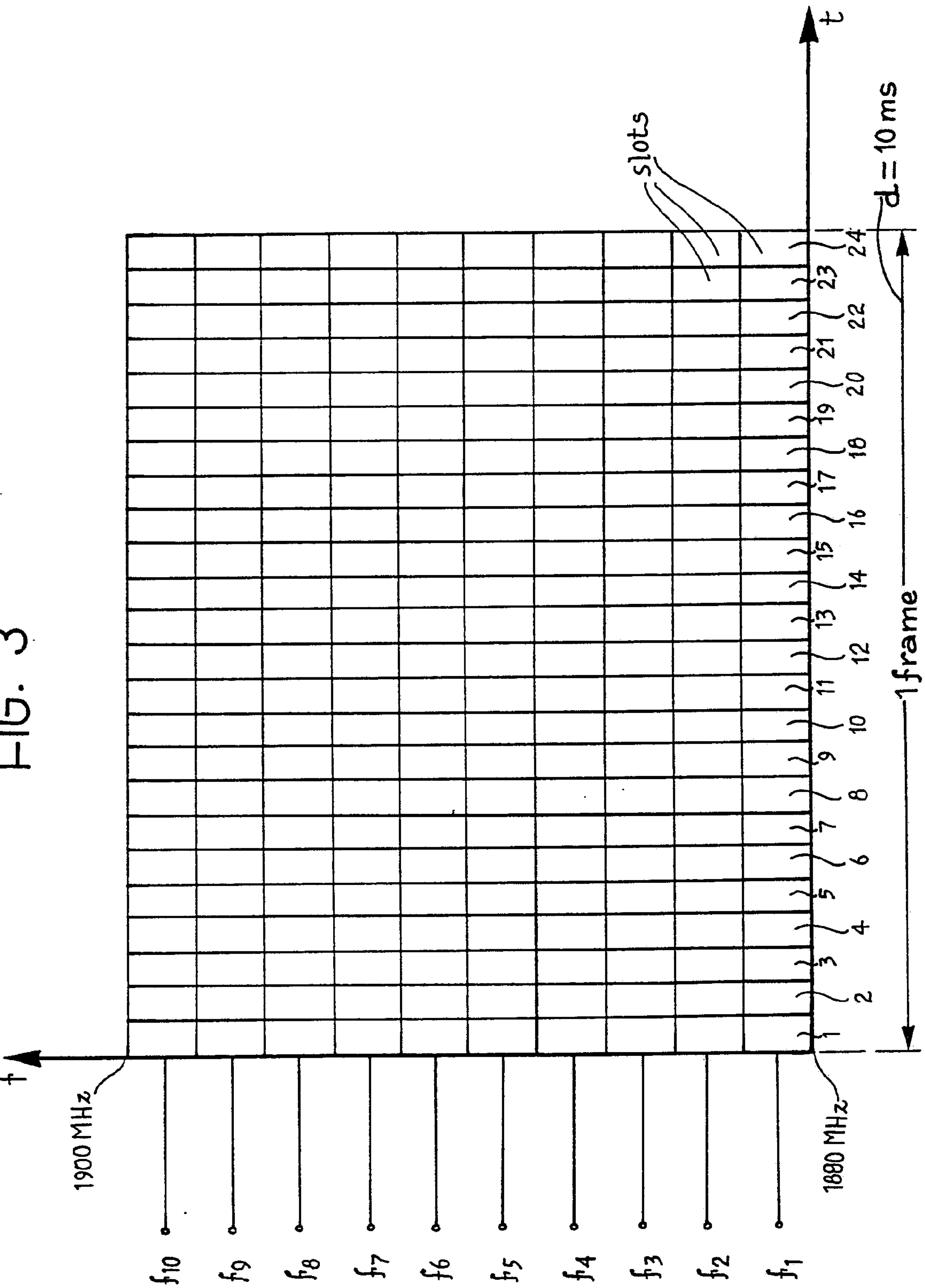
ASIC
59
MRM
RADIO TX/RX
60
A1
A2

FIG. 3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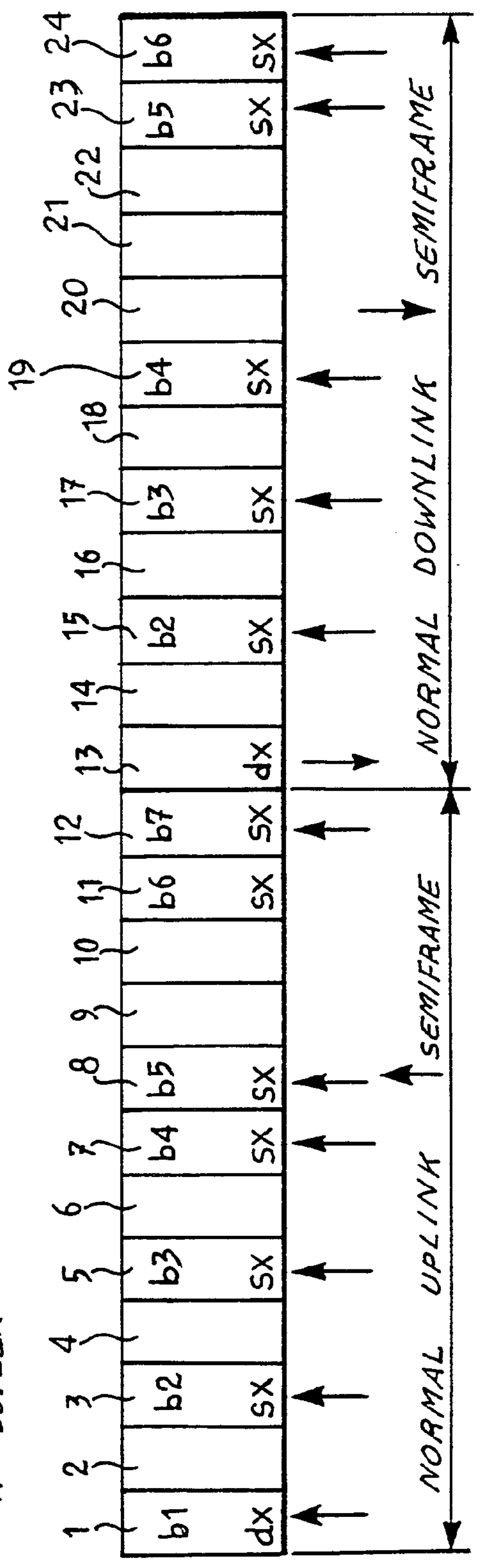
f
1900 MHz
1880 MHz
f_{10}
f_9
f_8
f_7
f_6
f_5
f_4
f_3
f_2
f_1

FIG. 4

Sx = DOUBLE SIMPLEX

dx = DUPLEX



INTERNATIONAL SEARCH REPORT

International Application No PCT/EP 92/02230

I. CLASSIFICATION OF SUBJECT MATTER (if several classification symbols apply, indicate all) [6]

According to International Patent Classification (IPC) or to both National Classification and IPC

IPC^5: H 04 B 7/24,H 04 H 3/00,H 04 L 12/44

II. FIELDS SEARCHED

Minimum Documentation Searched [7]

Classification System	Classification Symbols
IPC^5	H 04 B 1/00,H 04 B 9/00,H 04 J 3/00,H 04 L 11/00, H 04 L 12/00,H 04 N 5/00,H 04 Q 7/00

Documentation Searched other than Minimum Documentation to the Extent that such Documents are Included in the Fields Searched [8]

III. DOCUMENTS CONSIDERED TO BE RELEVANT [9]

Category *	Citation of Document, [11] with indication, where appropriate, of the relevant passages [12]	Relevant to Claim No. [13]
E P,A	US, A, 5 079 628 (TOMIKAWA) 07 January 1992 (07.01.92), see abstract; fig. 9; claims 1,10,15,18-20.	1,11, 12 2-10, 13-14
X A	EP, A2, 0 257 947 (ATT) 02 March 1988 (02.03.88), see abstract; fig. 1; claims 1-4,7,8.	1,11, 12 2-10, 13-14
A	US, A, 4 665 519 (KIRCHNER et al.) 12 May 1987 (12.05.87), see abstract; fig. 1-3; claims 1-8.	1-14
A	GB, A, 2 125 257 (PLESSEY) 29 February 1984 (29.02.84), see abstract; claims 1-3.	1-14

* Special categories of cited documents: [10]

"A" document defining the general state of the art which is not considered to be of particular relevance

"E" earlier document but published on or after the international filing date

"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art.

"&" document member of the same patent family

IV. CERTIFICATION

Date of the Actual Completion of the International Search	Date of Mailing of this International Search Report
28 December 1992	15 JAN 1993
International Searching Authority	**Signature of Authorized Officer**
EUROPEAN PATENT OFFICE	BLASL e.h.

International Application No PCT/EP 92/02230

III. DOCUMENTS CONSIDERED TO BE RELEVANT (CONTINUED FROM THE SECOND SHEET)

Category *	Citation of Document, [11] with indication, where appropriate, of the relevant passages	Relevant to Claim No.
A	DE, A1, 3 716 318 (BTS) 24 November 1988 (24.11.88), see abstract; claims 1,2,7-9.	1-14

ANHANG | ANNEX | ANNEXE

zum internationalen Recherchenbericht über die internationale Patentanmeldung Nr.

to the International Search Report to the International Patent Application No.

au rapport de recherche international relatif à la demande de brevet international n°

PCT/EP 92/02230 SAE 64827

In diesem Anhang sind die Mitglieder der Patentfamilien der im obengenannten internationalen Recherchenbericht angeführten Patentdokumente angegeben. Diese Angaben dienen nur zur Unterrichtung und erfolgen ohne Gewähr.

This Annex lists the patent family members relating to the patent documents cited in the above-mentioned international search report. The Office is in no way liable for these particulars which are given merely for the purpose of information.

La présente annexe indique les membres de la famille de brevets relatifs aux documents de brevets cités dans le rapport de recherche international visée ci-dessus. Les reseignements fournis sont donnés à titre indicatif et n'engagent pas la responsibilité de l'Office.

Im Recherchenbericht angeführtes Patentdokument / Patent document cited in search report / Document de brevet cité dans le rapport de recherche	Datum der Veröffentlichung / Publication date / Date de publication	Mitglied(er) der Patentfamilie / Patent family member(s) / Membre(s) de la famille de brevets	Datum der Veröffentlichung / Publication date / Date de publication
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		JP A2 1174124	10-07-89
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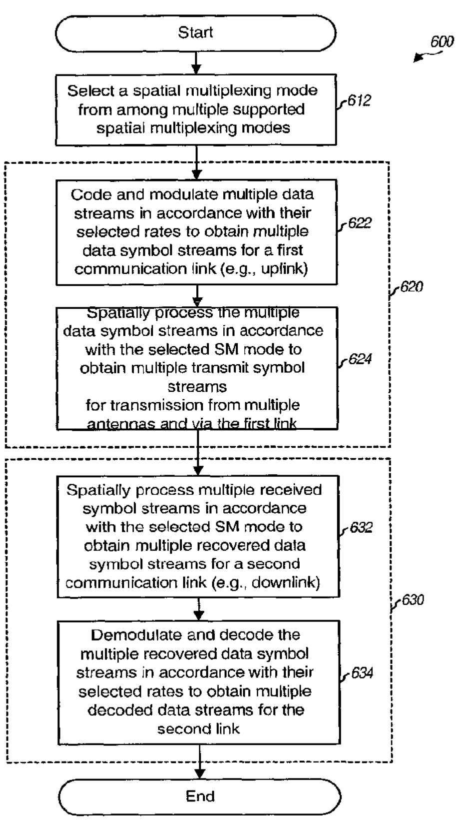
[Continued on next page]

(54) Title: MULTI-MODE TERMINAL IN A WIRELESS MIMO SYSTEM



(57) Abstract: A user terminal supports multiple spatial multiplexing (SM) modes such as a steered mode and a non-steered mode. For data transmission, multiple data streams are coded and modulated in accordance with their selected rates to obtain multiple data symbol streams. These streams are then spatially processed in accordance with a selected SM mode (*e.g.*, with a matrix of steering vectors for the steered mode and with the identity matrix for the non-steered mode) to obtain multiple transmit symbol streams for transmission from multiple antennas. For data reception, multiple received symbol streams are spatially processed in accordance with the selected SM mode (*e.g.*, with a matrix of eigenvectors for the steered mode and with a spatial filter matrix for the non-steered mode) to obtain multiple recovered data symbol streams. These streams are demodulated and decoded in accordance with their selected rates to obtain multiple decoded data streams.

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ES, FI, FR, GB, GR, HU, IE, IT, LU, MC, NL, PT, RO, SE, SI, SK, TR), OAPI patent (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:

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For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

MULTI-MODE TERMINAL IN A WIRELESS MIMO SYSTEM

Claim of Priority under 35 U.S.C. §119

[0001] The present Application for Patent claims priority to Provisional Application Serial No. 60/421,309, entitled "MIMO WLAN System," filed October 25, 2002, assigned to the assignee hereof, and expressly incorporated by reference herein.

BACKGROUND

Field

[0002] The present invention relates generally to communication, and more specifically to a user terminal in a multiple-input multiple-output (MIMO) communication system.

Background

[0003] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission and is denoted as an (N_T, N_R) system. A MIMO channel formed by the N_T transmit and N_R receive antennas may be decomposed into N_S spatial channels, where $N_S \leq \min\{N_T, N_R\}$. The N_S spatial channels may be used to transmit N_S independent data streams to achieve greater overall throughput. In general, spatial processing may or may not be performed at a transmitter and is normally performed at a receiver to simultaneously transmit and recover multiple data streams.

[0004] A conventional MIMO system typically uses a specific transmission scheme to simultaneously transmit multiple data streams. This transmission scheme may be selected based on a trade-off of various factors such as the requirements of the system, the amount of feedback from the receiver to the transmitter, the capabilities of the transmitter and receiver, and so on. The transmitter, receiver, and system are then designed to support and operate in accordance with the selected transmission scheme. This transmission scheme typically has favorable features as well as unfavorable ones, which can impact system performance.

[0005] There is therefore a need in the art for a user terminal capable of achieving improved performance.

SUMMARY

[0006] A user terminal that supports multiple spatial multiplexing (SM) modes for improved performance and greater flexibility is described herein. Spatial multiplexing refers to the transmission of multiple data streams simultaneously via multiple spatial channels of a MIMO channel. The multiple SM modes may include (1) a steered mode that transmits multiple data streams on orthogonal spatial channels and (2) a non-steered mode that transmits multiple data streams from multiple antennas.

[0007] The terminal selects an SM mode to use for data transmission from among the multiple supported SM modes. The SM mode selection may be based on various factors such as the calibration status of the terminal, the amount of data to send, the channel conditions, the capability of the other communicating entity, and so on. For data transmission, multiple data streams are coded and modulated in accordance with their selected rates to obtain multiple data symbol streams. These data symbol streams are then spatially processed in accordance with the selected SM mode to obtain multiple transmit symbol streams. The transmit spatial processing is with a matrix of steering vectors for the steered mode and with an identity matrix for the non-steered mode. The transmit symbol streams are transmitted from multiple antennas and via a first communication link (e.g., uplink).

[0008] For data reception, multiple received symbol streams for a second communication link (e.g., downlink) are spatially processed in accordance with the selected SM mode to obtain multiple recovered data symbol streams. The receive spatial processing may be based on the channel eigenvectors for the steered mode and with a spatial filter matrix for the non-steered mode. The spatial filter matrix may be derived based on various receiver spatial processing techniques, as described below. The recovered data symbol streams are then demodulated and decoded in accordance with their selected rates to obtain multiple decoded data streams for the second link. The terminal also transmits/receives pilots and the selected rates for each link.

[0009] Various aspects, embodiments, and features of the invention are described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[0010] FIG. 1 shows a MIMO system;

FIG. 2 shows spatial processing at a transmitter and receiver for the steered and non-steered modes;

FIGS. 3 and 4 show spatial processing at an access point and a user terminal for the steered and non-steered modes, respectively;

FIG. 5 shows a block diagram of the access point and user terminal; and

FIG. 6 shows a process for transmitting and receiving data in the MIMO system.

DETAILED DESCRIPTION

[0011] The word "exemplary" is used herein to mean "serving as an example, instance, or illustration." Any embodiment described herein as "exemplary" is not necessarily to be construed as preferred or advantageous over other embodiments.

[0012] **FIG. 1** shows a MIMO system 100 with access points (APs) and user terminals (UTs). For simplicity, only one access point 110 is shown in FIG. 1. An access point is generally a fixed station that communicates with the user terminals and may also be referred to as a base station or some other terminology. A system controller 130 couples to and provides coordination and control for the access points. A user terminal may be fixed or mobile and may also be referred to as a mobile station, a wireless device, or some other terminology. A user terminal may communicate with an access point, in which case the roles of access point and user terminal are established. A user terminal may also communicate peer-to-peer with another user terminal.

[0013] MIMO system 100 may be a time division duplex (TDD) system or a frequency division duplex (FDD) system. For a TDD system, the downlink and uplink share the same frequency band. For an FDD system, the downlink and uplink use different frequency bands. The downlink is the communication link from the access points to the user terminals, and the uplink is the communication link from the user terminals to the access points. MIMO system 100 may also utilize a single carrier or multiple carriers for data transmission.

[0014] Access point 110 and user terminal 120 each support multiple spatial multiplexing (SM) modes for improved performance and greater flexibility. A steered SM mode (or simply, a steered mode) can typically achieve better performance but can only be used if the transmitter has sufficient channel state information (CSI) to

orthogonalize the spatial channels of a MIMO channel via decomposition or some other technique. A non-steered SM mode (or simply, a non-steered mode) requires very little information to simultaneously transmit multiple data streams via the MIMO channel, but performance may not be quite as good as the steered mode. A suitable SM mode may be selected for use based on various factors, as described below.

[0015] Table 1 summarizes some key aspects of the steered and non-steered modes. Each SM mode has different capabilities and requirements.

[0016] For the steered mode, the transmitter transmits a pilot to allow the receiver to estimate the MIMO channel, and the receiver sends back sufficient channel state information to allow the transmitter to derive steering vectors. Either the transmitter or receiver decomposes the MIMO channel into eigenmodes, which may be viewed as orthogonal spatial channels. The receiver also sends back the rate to use for each eigenmode. The transmitter and receiver both perform spatial processing in order to transmit data on the eigenmodes, as described below.

[0017] For the non-steered mode, the transmitter transmits a pilot to allow the receiver to estimate the MIMO channel. The receiver sends back the rate to use for each spatial channel. The transmitter transmits data (e.g., from its antennas) without any spatial processing, and the receiver performs spatial processing to recover the transmitted data. The pilot transmission and spatial processing at the transmitter and receiver for the steered and non-steered modes are described below.

Table 1 – Requirements for Steered and Non-Steered Modes

	Steered Mode	**Non-Steered Mode**
Pilot	Transmitter transmits a pilot Receiver sends back channel state information used by transmitter to derive steering vectors	Transmitter transmits a pilot
Rate Feedback	Receiver sends back the rate for each eigenmode	Receiver sends back the rate for each spatial channel (e.g., each transmit antenna)
Spatial Processing	Transmitter performs spatial processing with matrix $\underline{\mathbf{V}}$ of steering vectors Receiver performs spatial processing with matrix $\underline{\mathbf{U}}$ of eigenvectors	Transmitter transmits data from each transmit antenna Receiver performs spatial processing with CCMI, MMSE, SIC, and so on (described below)

[0018] In the following description, a user terminal can be the transmitter and/or receiver, and an access point can likewise be the transmitter and/or receiver. Peer-to-peer communications can be supported using the same basic principles.

1. Steered Mode

[0019] A MIMO channel formed by N_T transmit antennas and N_R receive antennas may be characterized by an $N_R \times N_T$ channel response matrix $\underline{\mathbf{H}}$, which may be expressed as:

$$\underline{\mathbf{H}} = \begin{bmatrix} h_{1,1} & h_{1,2} & \cdots & h_{1,N_T} \\ h_{2,1} & h_{2,2} & \cdots & h_{2,N_T} \\ \vdots & \vdots & \ddots & \vdots \\ h_{N_R,1} & h_{N_R,2} & \cdots & h_{N_R,N_T} \end{bmatrix}, \qquad \text{Eq (1)}$$

where entry $h_{i,j}$, for $i = 1 \ldots N_R$ and $j = 1 \ldots N_T$, is the coupling (i.e., complex gain) between transmit antenna j and receive antenna i. For simplicity, the MIMO channel is assumed to be full rank with $N_S \le N_T \le N_R$.

[0020] Singular value decomposition may be performed on $\underline{\mathbf{H}}$ to obtain N_S eigenmodes of $\underline{\mathbf{H}}$, as follows:

$$\underline{\mathbf{H}} = \underline{\mathbf{U}}\underline{\Sigma}\underline{\mathbf{V}}^H \ , \qquad \text{Eq (2)}$$

where $\underline{\mathbf{U}}$ is an $(N_R \times N_R)$ unitary matrix of left eigenvectors of $\underline{\mathbf{H}}$;

$\underline{\Sigma}$ is an $(N_R \times N_T)$ diagonal matrix of singular values of $\underline{\mathbf{H}}$;

$\underline{\mathbf{V}}$ is an $(N_T \times N_T)$ unitary matrix of right eigenvectors of $\underline{\mathbf{H}}$; and

"^{H}" denotes the conjugate transpose.

A unitary matrix $\underline{\mathbf{M}}$ is characterized by the property $\underline{\mathbf{M}}^H\underline{\mathbf{M}} = \underline{\mathbf{I}}$, where $\underline{\mathbf{I}}$ is the identity matrix. The columns of a unitary matrix are orthogonal to one another.

[0021] The right eigenvectors of $\underline{\mathbf{H}}$ are also referred to as steering vectors and may be used for spatial processing by the transmitter to transmit data on the N_S eigenmodes of $\underline{\mathbf{H}}$. The left eigenvectors of $\underline{\mathbf{H}}$ may be used for spatial processing by the receiver to recover the data transmitted on the N_S eigenmodes. The eigenmodes may be viewed as orthogonal spatial channels obtained through decomposition. The diagonal entries of $\underline{\Sigma}$ are the singular values of $\underline{\mathbf{H}}$, which represent the channel gains for the N_S eigenmodes.

[0022] In a practical system, only an estimate of $\underline{\mathbf{H}}$ can be obtained, and only estimates of $\underline{\mathbf{V}}$, $\underline{\Sigma}$ and $\underline{\mathbf{U}}$ can be derived. The N_S spatial channels are also typically not completely orthogonal to one another due to various reasons such as an imperfect channel estimate. For simplicity, the description herein assumes channel estimation and decomposition without errors. Furthermore, the term "eigenmode" covers the case where an attempt is made to orthogonalize the spatial channels using decomposition, even though the attempt may not be fully successful due to, for example, an imperfect channel estimate.

[0023] Table 2 summarizes the spatial processing at the transmitter and the receiver for the steered mode. In Table 2, $\underline{\mathbf{s}}$ is a vector with N_S data symbols to be transmitted on the N_S eigenmodes of $\underline{\mathbf{H}}$, $\underline{\mathbf{x}}_{st}$ is a vector with N_T transmit symbols to be sent from the N_T transmit antennas, $\underline{\mathbf{r}}_{st}$ is a vector with N_R received symbols obtained from the N_R receive antennas, $\hat{\underline{\mathbf{s}}}_{st}$ is a vector with N_S recovered data symbols (i.e., $\hat{\underline{\mathbf{s}}}_{st}$ is an estimate of $\underline{\mathbf{s}}$), and the subscript "*st*" denotes the steered mode. As used herein, a "data symbol" refers to a modulation symbol for data, and a "pilot symbol" refers to a modulation symbol for pilot.

Table 2 – Spatial Processing for Steered Mode

Transmitter	Receiver
$\underline{\mathbf{x}}_{st} = \underline{\mathbf{V}}\underline{\mathbf{s}}$	$\hat{\underline{\mathbf{s}}}_{st} = \underline{\Sigma}^{-1}\underline{\mathbf{U}}^{H}\underline{\mathbf{r}}_{st}$

[0024] Eigenvalue decomposition may also be performed on a correlation matrix of $\underline{\mathbf{H}}$, which is $\underline{\mathbf{R}} = \underline{\mathbf{H}}^{H}\underline{\mathbf{H}}$, as follows:

$$\underline{\mathbf{R}} = \underline{\mathbf{H}}^{H}\underline{\mathbf{H}} = \underline{\mathbf{V}}\underline{\Lambda}\underline{\mathbf{V}}^{H} \,, \qquad \text{Eq (3)}$$

where $\underline{\Lambda}$ is a diagonal matrix of eigenvalues, which are the squares of the singular values in $\underline{\Sigma}$. The transmitter can perform spatial processing with $\underline{\mathbf{V}}$ to obtain $\underline{\mathbf{x}}_{st}$, and the receiver can perform spatial processing with $\underline{\mathbf{V}}^{H}\underline{\mathbf{H}}^{H}$ to obtain $\hat{\underline{\mathbf{s}}}_{st}$.

2. Non-Steered Mode

[0025] For the non-steered mode, the transmitter can transmit one data symbol stream from each transmit antenna. A spatial channel for this mode can correspond to one transmit antenna. The receiver performs spatial processing to separate out and recover the transmitted data symbol streams. The receiver can use various receiver processing techniques such as a channel correlation matrix inversion (CCMI) technique (which is also known as a zero-forcing technique), a minimum mean square error (MMSE) technique, a successive interference cancellation (SIC) technique, and so on

[0026] Table 3 summarizes the spatial processing at the transmitter and the receiver for the non-steered mode. In Table 3, $\underline{\mathbf{x}}_{ns}$ is a vector with N_T data symbols to be sent from the N_T transmit antennas, $\underline{\mathbf{r}}_{ns}$ is a vector with N_R received symbols obtained from the N_R receive antennas, $\underline{\mathbf{M}}_{ccmi}$ is a spatial filter matrix for the CCMI technique, $\underline{\mathbf{M}}_{mmse}$ is a spatial filter matrix for the MMSE technique, $\underline{\mathbf{D}}_{mmse}$ is a diagonal matrix for the MMSE technique (which contains the diagonal elements of $\underline{\mathbf{M}}_{mmse}\underline{\mathbf{H}}$), and the subscript "*ns*" denotes the non-steered mode.

Table 3 – Spatial Processing for Non-Steered Mode

Transmitter	Receiver	where	
$\underline{\mathbf{x}}_{ns} = \underline{\mathbf{s}}$	$\hat{\underline{\mathbf{s}}}_{ccmi} = \underline{\mathbf{M}}_{ccmi}\underline{\mathbf{r}}_{ns}$	$\underline{\mathbf{M}}_{ccmi} = [\underline{\mathbf{H}}^H \underline{\mathbf{H}}]^{-1}\underline{\mathbf{H}}^H$	CCMI
	$\hat{\underline{\mathbf{s}}}_{mmse} = \underline{\mathbf{D}}_{mmse}^{-1}\underline{\mathbf{M}}_{mmse}\underline{\mathbf{r}}_{ns}$	$\underline{\mathbf{M}}_{mmse} = \underline{\mathbf{H}}^H [\underline{\mathbf{H}}\underline{\mathbf{H}}^H + \sigma^2\underline{\mathbf{I}}]^{-1}$ and $\underline{\mathbf{D}}_{mmse} = \text{diag } [\underline{\mathbf{M}}_{mmse}\underline{\mathbf{H}}]$	MMSE
	$\hat{\underline{\mathbf{s}}}_{sic}^{\lambda} = \underline{\mathbf{M}}_{sic}^{\lambda}\underline{\mathbf{r}}_{sic}^{\lambda}$	$\underline{\mathbf{M}}_{sic}^{\lambda} = \underline{\mathbf{M}}_{ccmi}^{\lambda}$ or $(\underline{\mathbf{D}}_{mmse}^{\lambda})^{-1}\underline{\mathbf{M}}_{mmse}^{\lambda}$	SIC

For simplicity, the MIMO channel noise $\underline{\mathbf{n}}$ is assumed to be additive white Gaussian noise (AWGN) with zero mean, a variance of σ^2, and an autocovariance matrix of $\underline{\varphi}_{nn} = E[\underline{\mathbf{n}}\underline{\mathbf{n}}^H] = \sigma^2\underline{\mathbf{I}}$.

[0027] For the SIC technique, the receiver processes the N_R received symbol streams in N_S successive stages to recover one data symbol stream in each stage. For each stage λ, where $\lambda = 1 \ldots N_S$, the receiver initially performs spatial processing on N_R input symbol streams for stage λ using the CCMI, MMSE, or some other technique and obtains one recovered data symbol stream. The N_R received symbol streams are the N_R input symbol streams for stage 1. The receiver further processes (e.g., demodulates, deinterleaves, and decodes) the recovered data symbol stream for stage λ to obtain a decoded data stream, estimates the interference this stream causes to the other data symbol streams not yet recovered, and cancels the estimated interference from the N_R input symbol streams for stage λ to obtain N_R input symbol streams for stage $\lambda+1$. The receiver then repeats the same processing on the N_R input symbol streams for stage $\lambda+1$ to recover another data symbol stream.

[0028] For each stage λ, the SIC receiver derives a spatial filter matrix $\underline{\mathbf{M}}_{sic}^{\lambda}$ for that stage based on a reduced channel response matrix $\underline{\mathbf{H}}^{\lambda}$ and using the CCMI, MMSE, or some other technique. The reduced matrix $\underline{\mathbf{H}}^{\lambda}$ is obtained by removing $\lambda-1$ columns in the original matrix $\underline{\mathbf{H}}$ corresponding to the $\lambda-1$ data symbol streams already recovered. The matrix $\underline{\mathbf{M}}_{sic}^{\lambda}$ has dimensionality of $(N_T - \lambda + 1) \times N_R$. Since $\underline{\mathbf{H}}^{\lambda}$ is different for each stage, $\underline{\mathbf{M}}_{sic}^{\lambda}$ is also different for each stage.

[0029] The receiver may also use other receiver spatial processing techniques to recover the transmitted data symbol streams.

[0030] FIG. 2 shows the spatial processing at the transmitter and receiver for the steered and non-steered modes. At the transmitter, the data vector $\underline{\mathbf{s}}$ is multiplied with either the matrix $\underline{\mathbf{V}}$ for the steered mode or the identity matrix $\underline{\mathbf{I}}$ for the non-steered mode by a unit 220 to obtain the transmit symbol vector $\underline{\mathbf{x}}$. At the receiver, the received symbol vector $\underline{\mathbf{r}}$ is multiplied with either the matrix $\underline{\mathbf{U}}^H$ for the steered mode or the spatial filter matrix $\underline{\mathbf{M}}$ for the non-steered mode by a unit 260 to obtain a detected symbol vector $\tilde{\underline{\mathbf{s}}}$, which is an unnormalized estimate of $\underline{\mathbf{s}}$. The matrix $\underline{\mathbf{M}}$ may be derived based on the CCMI, MMSE, or some other technique. The vector $\tilde{\underline{\mathbf{s}}}$ is further scaled with either the diagonal matrix $\underline{\Sigma}^{-1}$ for the steered mode or a diagonal matrix $\underline{\mathbf{D}}^{-1}$ for the non-steered mode to obtain the recovered data symbol vector $\hat{\underline{\mathbf{s}}}$, where $\underline{\mathbf{D}}^{-1} = \underline{\mathbf{I}}$ for the CCMI technique and $\underline{\mathbf{D}}^{-1} = \underline{\mathbf{D}}_{mmse}^{-1}$ for the MMSE technique.

3. Overhead for Steered and Non-Steered Nodes

[0031] The steered and non-steered modes have different pilot and overhead requirements, as shown in

[0032] Table 1 and described below.

A. Pilot Transmission

[0033] For both the steered and non-steered modes, the transmitter can transmit a MIMO pilot (which is an unsteered pilot) to allow the receiver to estimate the MIMO channel and obtain the matrix $\underline{\mathbf{H}}$. The MIMO pilot comprises N_T orthogonal pilot transmissions sent from N_T transmit antennas, where orthogonality may be achieved in time, frequency, code, or a combination thereof. For code orthogonality, the N_T pilot transmissions can be sent simultaneously from the N_T transmit antennas, with the pilot transmission from each antenna being "covered" with a different orthogonal (e.g., Walsh) sequence. The receiver "decovers" the received pilot symbols for each receive antenna i with the same N_T orthogonal sequences used by the transmitter to obtain estimates of the complex channel gain between receive antenna i and each of the N_T transmit antennas. The covering at the transmitter and the decovering at the receiver are performed in similar manner as for a Code Division Multiple Access (CDMA) system. For frequency orthogonality, the N_T pilot transmissions for the N_T transmit antennas can be sent simultaneously on different subbands of the overall system bandwidth. For time

orthogonality, the N_T pilot transmissions for the N_T transmit antennas can be sent in different time slots. In any case, the orthogonality among the N_T pilot transmissions allows the receiver to distinguish the pilot transmission from each transmit antenna.

[0034] For the steered mode, the receiver sends back sufficient channel state information to allow the transmitter to derive the steering vectors. The receiver may send this information in a direct form (e.g., by sending the entries of $\underline{\mathbf{V}}$) or in an indirect form (e.g., by transmitting a steered or unsteered pilot).

B. <u>Rate Selection/Control</u>

[0035] The receiver can estimate the received signal-to-noise-and-interference ratio (SNR) for each spatial channel, which can correspond to an eigenmode for the steered mode or a transmit antenna for the non-steered mode. The received SNR is dependent on the SM mode and spatial processing technique used by the transmitter and receiver.

[0036] Table 4 summarizes the received SNR for the steered and non-steered modes. In Table 4, P_m is the transmit power used for spatial channel m, σ^2 is the noise variance, σ_m is the singular value for eigenmode m (i.e., the m-th diagonal element of $\underline{\Sigma}$), r_{mm} is the m-th diagonal element of $\underline{\mathbf{R}}$ (which is $\underline{\mathbf{R}} = \underline{\mathbf{H}}^H \underline{\mathbf{H}}$), q_{mm} is the m-th diagonal element of $\underline{\mathbf{Q}}$, and γ_m is the SNR for spatial channel m. The received SNRs for the SIC technique are dependent on the spatial processing technique (e.g., CCMI or MMSE) and the order in which the data streams are recovered. An operating SNR can be defined as being equal to the received SNR plus an SNR back-off factor. The SNR back-off factor can be set to a positive value to account for estimation error, SNR fluctuation over time, and so on, but may also be set to zero.

Table 4 – Received SNR

Steered Mode	Non-Steered Mode	
	CCMI	MMSE
$\gamma_{st,m} = \frac{P_m \cdot \sigma_m^2}{\sigma^2}$	$\gamma_{ccmi,m} = \frac{P_m}{r_{mm} \cdot \sigma^2}$	$\gamma_{mmse,m} = \frac{q_{mm}}{1-q_{mm}} P_m$

[0037] The MIMO system may support a set of rates. Each non-zero rate is associated with a particular data rate or spectral efficiency, a particular coding scheme, a particular modulation scheme, and a particular SNR required to achieve a target level of performance (e.g., one percent packet error rate (PER)). The required SNR for each rate may be determined by computer simulation, empirical measurement, and so on, and with an assumption of an AWGN channel. A look-up table (LUT) can store the rates supported by the system and their required SNRs. For each spatial channel, the highest rate in the look-up table with a required SNR that is equal to or less than the operating SNR of the spatial channel is selected as the rate to use for the spatial channel.

[0038] Closed-loop rate control may be used for each spatial channel or a combination of spatial channels. The receiver can estimate the received SNR for each spatial channel, select the proper rate for the spatial channel, and send back the selected rate. The transmitter can transmit each data symbol stream at the selected rate.

C. Mode Selection

[0039] User terminal 120 can use either the steered or non-steered mode at any given moment for communication. The mode selection may be made based on various factors such as the following.

[0040] Overhead – The steered mode requires more overhead than the non-steered mode. For the steered mode, the receiver needs to send back sufficient channel state information as well as the rates for the N_S eigenmodes. In some instances, the additional CSI overhead cannot be supported or is not justified. For the non-steered mode, the receiver only needs to send back the rates for the spatial channels, which is much less overhead.

[0041] Amount of Data – The steered mode is generally more efficient but also requires more setup steps (e.g., channel estimation, singular value decomposition, and CSI

feedback). If only a small amount of data needs to be sent, then it may be quicker and more efficient to transmit this data using the non-steered mode.

[0042] Capability – A user terminal may communicate peer-to-peer with another user terminal that supports only one mode (e.g., either the steered or non-steered mode). In this case, the two terminals can communicate using a common mode supported by both user terminals.

[0043] Channel Conditions – The steered mode may be more easily supported for static channels, slow varying channels, and channels with a strong line-of-site component (e.g., a Rician channel).

[0044] Receiver SNR – The steered mode provides better performance in low SNR conditions. A user terminal may elect to use steered mode when the SNR drops below some threshold.

[0045] Calibration Status – The steered mode may be selected for use if the transmitter and receiver are "calibrated" such that the downlink and uplink channel responses are reciprocal of one another. Reciprocal downlink and uplink can simplify the pilot transmission and spatial processing for both the transmitter and receiver for the steered mode, as described below.

[0046] A user terminal that is not mobile and is communicating with the same access point may use the steered mode much of the time. A user terminal that is mobile and communicating with different entities (e.g., different access points and/or other user terminals) may use the non-steered mode, until such time that it is more advantageous to use the steered mode. A user terminal may also switch between the steered and non-steered modes, as appropriate. For example, a user terminal may use the non-steered mode for small data bursts (or short data sessions) and at the start of long data bursts (or long data sessions), and may use the steered mode for the remaining portion of the long data bursts. As another example, a user terminal may use the steered mode for relatively static channel conditions and may use the non-steered mode when the channel conditions change more rapidly.

4. TDD MIMO System

[0047] A multi-mode user terminal for an exemplary MIMO wireless local area network (WLAN) system is described below. The MIMO WLAN system utilizes orthogonal frequency division multiplexing (OFDM), which is a multi-carrier modulation technique

that effectively partitions the overall system bandwidth into multiple (N_F) orthogonal subbands. With OFDM, each subband is associated with a respective carrier that may be modulated with data.

[0048] The exemplary MIMO WLAN system is a TDD system. A high degree of correlation normally exists between the downlink and uplink channel responses for the TDD system since these links share the same frequency band. However, the responses of the transmit/receive chains at the access point are typically not the same as the responses of the transmit/receive chains at the user terminal. The differences can be determined and accounted for via calibration. The overall downlink and uplink channel responses may then be assumed to be reciprocal (i.e., transpose) of each other. The channel estimation and spatial processing for the steered mode can be simplified with reciprocal downlink and uplink.

[0049] **FIG. 3** shows the transmit/receive chains at access point 110 and user terminal 120. At access point 110, transmit chain 324 and receive chain 334 are modeled by matrices $\underline{\mathbf{T}}_{\mathrm{ap}}(k)$ and $\underline{\mathbf{R}}_{\mathrm{ap}}(k)$, respectively, for each subband k. At user terminal 120, transmit chain 364 and receive chain 354 are modeled by matrices $\underline{\mathbf{T}}_{\mathrm{ut}}(k)$ and $\underline{\mathbf{R}}_{\mathrm{ut}}(k)$, respectively, for each subband k.

[0050] Table 5 summarizes the calibration and singular value decomposition for the downlink and uplink in the TDD MIMO WLAN system. The "effective" downlink and uplink channel responses, $\underline{\mathbf{H}}_{\mathrm{edn}}(k)$ and $\underline{\mathbf{H}}_{\mathrm{eup}}(k)$, include the responses of the appropriate transmit and receive chains. Diagonal correction matrices $\underline{\mathbf{K}}_{\mathrm{ap}}(k)$ and $\underline{\mathbf{K}}_{\mathrm{ut}}(k)$ are obtained by performing calibration with MIMO pilots transmitted by both the access point and user terminal. The "calibrated" downlink and uplink channel responses, $\underline{\mathbf{H}}_{\mathrm{cdn}}(k)$ and $\underline{\mathbf{H}}_{\mathrm{cup}}(k)$, include the correction matrices and are reciprocal of one another (i.e., $\underline{\mathbf{H}}_{\mathrm{cup}}(k) = \underline{\mathbf{H}}_{\mathrm{cdn}}^{T}(k)$, where "^{T}" denotes the transpose).

Table 5 – Channel Responses for TDD MIMO WLAN System

	Downlink	Uplink
Effective Channel Response	$\underline{\mathbf{H}}_{\text{edn}}(k)=\underline{\mathbf{R}}_{\text{ut}}(k)\underline{\mathbf{H}}(k)\underline{\mathbf{T}}_{\text{ap}}(k)$	$\underline{\mathbf{H}}_{\text{eup}}(k)=\underline{\mathbf{R}}_{\text{ap}}(k)\underline{\mathbf{H}}^{T}(k)\underline{\mathbf{T}}_{\text{ut}}(k)$
Correction Matrix	$\underline{\mathbf{K}}_{\text{ap}}(k)=\underline{\mathbf{T}}_{\text{ap}}^{-1}(k)\underline{\mathbf{R}}_{\text{ap}}(k)$	$\underline{\mathbf{K}}_{\text{ut}}(k)=\underline{\mathbf{T}}_{\text{ut}}^{-1}(k)\underline{\mathbf{R}}_{\text{ut}}(k)$
Calibrated Channel Response	$\underline{\mathbf{H}}_{\text{cdn}}(k)=\underline{\mathbf{H}}_{\text{edn}}(k)\underline{\mathbf{K}}_{\text{ap}}(k)$	$\underline{\mathbf{H}}_{\text{cup}}(k)=\underline{\mathbf{H}}_{\text{eup}}(k)\underline{\mathbf{K}}_{\text{ut}}(k)$
Singular Value Decomposition	$\underline{\mathbf{H}}_{\text{cdn}}(k)=\underline{\mathbf{V}}_{\text{ut}}^{*}(k)\underline{\Sigma}(k)\underline{\mathbf{U}}_{\text{ap}}^{T}(k)$	$\underline{\mathbf{H}}_{\text{cup}}(k)=\underline{\mathbf{U}}_{\text{ap}}(k)\underline{\Sigma}(k)\underline{\mathbf{V}}_{\text{ut}}^{H}(k)$

[0051] Because $\underline{\mathbf{H}}_{\text{cup}}(k)$ and $\underline{\mathbf{H}}_{\text{cdn}}(k)$ are reciprocal, the matrices $\underline{\mathbf{V}}_{\text{ut}}^{*}(k)$ and $\underline{\mathbf{U}}_{\text{ap}}^{*}(k)$ of left and right eigenvectors of $\underline{\mathbf{H}}_{\text{cdn}}(k)$ are the complex conjugate of the matrices $\underline{\mathbf{V}}_{\text{ut}}(k)$ and $\underline{\mathbf{U}}_{\text{ap}}(k)$ of right and left eigenvectors of $\underline{\mathbf{H}}_{\text{cup}}(k)$. The matrix $\underline{\mathbf{U}}_{\text{ap}}(k)$ can be used by access point 110 for both transmit and receive spatial processing. The matrix $\underline{\mathbf{V}}_{\text{ut}}(k)$ can be used by user terminal 120 for both transmit and receive spatial processing.

[0052] Singular value decomposition may be performed independently for each of the N_F subbands. For each subband, the singular values in $\underline{\Sigma}(k)$ may be ordered from largest to smallest, and the eigenvectors in $\underline{\mathbf{V}}(k)$ and $\underline{\mathbf{U}}(k)$ may be ordered correspondingly. A "wideband" eigenmode may be defined as the set of same-order eigenmodes for all N_F subbands after the ordering. The decomposition only needs to be performed by either user terminal 120 or access point 110. If performed by user terminal 120, then the matrices $\underline{\mathbf{U}}_{\text{ap}}(k)$, for $k=1 \ldots N_F$, may be provided to access point 110 in either a direct form (e.g., by sending entries of $\underline{\mathbf{U}}_{\text{ap}}(k)$) or an indirect form (e.g., by transmitting a steered pilot).

[0053] Table 6 summarizes the spatial processing at access point 110 and user terminal 120 for data transmission and reception on the downlink and uplink in the TDD MIMO WLAN system for the steered mode. In Table 6, the subscript "up" denotes the uplink, and the subscript "dn" denotes the downlink.

Table 6 – Spatial Processing for Steered Mode in TDD MIMO WLAN System

	Downlink	Uplink
Access Point	Transmit: $\underline{\mathbf{x}}_{dn}(k) = \underline{\mathbf{K}}_{ap}(k)\underline{\mathbf{U}}_{ap}^{*}(k)\underline{\mathbf{s}}_{dn}(k)$	Receive: $\hat{\underline{\mathbf{s}}}_{up}(k) = \underline{\Sigma}^{-1}(k)\underline{\mathbf{U}}_{ap}^{H}(k)\underline{\mathbf{r}}_{up}(k)$
User Terminal	Receive: $\hat{\underline{\mathbf{s}}}_{dn}(k) = \underline{\Sigma}^{-1}(k)\underline{\mathbf{V}}_{ut}^{T}(k)\underline{\mathbf{r}}_{dn}(k)$	Transmit: $\underline{\mathbf{x}}_{up}(k) = \underline{\mathbf{K}}_{ut}(k)\underline{\mathbf{V}}_{ut}(k)\underline{\mathbf{s}}_{up}(k)$

[0054] For the steered mode, the access point can transmit a MIMO pilot on the downlink. The user terminal can estimate the calibrated downlink channel based on the MIMO pilot, perform singular value decomposition, and transmit a steered pilot on the uplink using the matrix $\underline{\mathbf{V}}_{ut}(k)$. A steered pilot is a pilot transmitted on the eigenmodes using the same steering vectors that are used for data transmission on the eigenmodes. The access point can directly estimate the matrix $\underline{\mathbf{U}}_{ap}(k)$ based on the uplink steered pilot. Pilots may also be transmitted in other manners for the steered mode. For example, the user terminal can transmit the MIMO pilot, and the access point can transmit the steered pilot. As another example, the access point and user terminal can both transmit MIMO pilots.

[0055] For the non-steered mode, the transmitter (either the access point or user terminal) can transmit a MIMO pilot along with the data transmission. The receiver performs spatial processing (e.g., with CCMI, MMSE, SIC, or some other technique) to recover the data symbol streams, as described above.

[0056] Table 7 summarizes an embodiment of the pilot transmission and spatial processing for the steered and non-steered modes for the TDD MIMO WLAN system.

Table 7 – Data Transmission in TDD MIMO WLAN System

	Steered Mode	Non-Steered Mode
Calibration	Calibration is performed	Calibration is not required
Downlink Data Transmission	AP transmits a MIMO pilot UT transmits a steered pilot	AP transmits a MIMO pilot
	UT sends the rate for each downlink wideband eigenmode	UT sends the rate for each downlink wideband spatial channel
	AP transmits data with $\underline{\mathbf{U}}_{\mathrm{ap}}(k)$ UT receives data with $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$	AP transmits data from each antenna UT receives data with CCMI, MMSE, SIC, and so on
Uplink Data Transmission	AP transmits a MIMO pilot UT transmits a steered pilot	UT transmits a MIMO pilot
	AP sends the rate for each uplink wideband eigenmode	AP sends the rate for each uplink wideband spatial channel
	UT transmits data with $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$ AP receives data with $\underline{\mathbf{U}}_{\mathrm{ap}}(k)$	UT transmits data from each antenna AP receives data with CCMI, MMSE, SIC, and so on

[0057] For both the steered and non-steered modes, the receiver (either the access point or user terminal) can estimate the average received SNR for each wideband spatial channel, for example, by averaging the received SNRs (in dB) for the N_F subbands of the wideband spatial channel. A wideband spatial channel can correspond to a wideband eigenmode for the steered mode or a transmit antenna for the non-steered mode. The receiver then computes an operating SNR for each wideband spatial channel as the sum of the average received SNR plus the SNR back-off factor. The receiver then selects the rate for each wideband spatial channel based on the operating SNR and the look-up table of supported rates and their required SNRs.

[0058] **FIG. 3** shows the spatial processing at access point 110 and user terminal 120 for downlink and uplink data transmission for the steered mode in the MIMO WLAN system. For the downlink, at access point 110, the data symbol vector $\underline{\mathbf{s}}_{\mathrm{dn}}(k)$ is multiplied with the matrix $\underline{\mathbf{U}}^{*}_{\mathrm{ap}}(k)$ by a unit 320 and further scaled with the correction matrix $\underline{\mathbf{K}}_{\mathrm{ap}}(k)$ by a unit 322 to obtain the transmit symbol vector $\underline{\mathbf{x}}_{\mathrm{dn}}(k)$ for the

downlink. At user terminal 120, the received symbol vector $\underline{\mathbf{r}}_{\mathrm{dn}}(k)$ is multiplied with the matrix $\underline{\mathbf{V}}_{\mathrm{ut}}^{T}(k)$ by a unit 360 and further scaled with the matrix $\underline{\Sigma}^{-1}(k)$ by a unit 362 to obtain the recovered data symbol vector $\hat{\underline{\mathbf{s}}}_{\mathrm{dn}}(k)$ for the downlink.

[0059] For the uplink, at user terminal 120, the data symbol vector $\underline{\mathbf{s}}_{\mathrm{up}}(k)$ is multiplied with the matrix $\underline{\mathbf{V}}_{\mathrm{ut}}(k)$ by a unit 390 and further scaled with the correction matrix $\underline{\mathbf{K}}_{\mathrm{ut}}(k)$ by a unit 392 to obtain the transmit symbol vector $\underline{\mathbf{x}}_{\mathrm{up}}(k)$ for the uplink. At access point 110, the received symbol vector $\underline{\mathbf{r}}_{\mathrm{up}}(k)$ is multiplied with the matrix $\underline{\mathbf{U}}_{\mathrm{ap}}^{H}(k)$ by a unit 340 and further scaled with the matrix $\underline{\Sigma}^{-1}(k)$ by a unit 342 to obtain the recovered data symbol vector $\hat{\underline{\mathbf{s}}}_{\mathrm{up}}(k)$ for the uplink.

[0060] **FIG. 4** shows the spatial processing at access point 110 and user terminal 120 for downlink and uplink data transmission for the non-steered mode in the MIMO WLAN system. For the downlink, at access point 110, the data symbol vector $\underline{\mathbf{s}}_{\mathrm{dn}}(k)$ is multiplied with the identity matrix $\underline{\mathbf{I}}$ by a unit 420 to obtain the transmit symbol vector $\underline{\mathbf{x}}_{\mathrm{dn}}(k)$ for the downlink. At user terminal 120, the received symbol vector $\underline{\mathbf{r}}_{\mathrm{dn}}(k)$ is multiplied with a spatial filter matrix $\underline{\mathbf{M}}_{\mathrm{ut}}(k)$ by a unit 460 and further scaled with a diagonal matrix $\underline{\mathbf{D}}_{\mathrm{ut}}^{-1}(k)$ by a unit 462 to obtain the recovered data symbol vector $\hat{\underline{\mathbf{s}}}_{\mathrm{dn}}(k)$ for the downlink. The matrices $\underline{\mathbf{M}}_{\mathrm{ut}}(k)$ and $\underline{\mathbf{D}}_{\mathrm{ut}}^{-1}(k)$ are derived based on the effective downlink channel response matrix $\underline{\mathbf{H}}_{\mathrm{edn}}(k)$ and using the CCMI, MMSE, SIC, or some other technique.

[0061] For the uplink, at user terminal 120, the data symbol vector $\underline{\mathbf{s}}_{\mathrm{up}}(k)$ is multiplied with the identity matrix $\underline{\mathbf{I}}$ by a unit 490 to obtain the transmit symbol vector $\underline{\mathbf{x}}_{\mathrm{up}}(k)$ for the uplink. At access point 110, the received symbol vector $\underline{\mathbf{r}}_{\mathrm{up}}(k)$ is multiplied with a spatial filter matrix $\underline{\mathbf{M}}_{\mathrm{ap}}(k)$ by a unit 440 and further scaled with a diagonal matrix $\underline{\mathbf{D}}_{\mathrm{ap}}^{-1}(k)$ by a unit 442 to obtain the recovered data symbol vector $\hat{\underline{\mathbf{s}}}_{\mathrm{up}}(k)$ for the uplink. The matrices $\underline{\mathbf{M}}_{\mathrm{ap}}(k)$ and $\underline{\mathbf{D}}_{\mathrm{ap}}^{-1}(k)$ are derived based on the effective uplink channel response matrix $\underline{\mathbf{H}}_{\mathrm{eup}}(k)$ and using the CCMI, MMSE, SIC, or some other technique.

[0062] **FIG. 5** shows a block diagram of access point 110 and user terminal 120. On the downlink, at access point 110, a transmit (TX) data processor 514 receives traffic

data from a data source 512 and control data from a controller 530. TX data processor 514 processes (e.g., encodes, interleaves, and symbol maps) each of N_S data streams based on the coding and modulation schemes corresponding to the rate selected for the stream to obtain a data symbol stream. A TX spatial processor 520 receives N_S data symbol streams from TX data processor 514, performs spatial processing (as required) on the data symbols, multiplexes in pilot symbols, and provides N_{ap} transmit symbol streams for the N_{ap} antennas. The processing by TX spatial processor 520 is dependent on whether the steered or non-steered mode is selected for use and may be performed as described above. Each transmitter unit (TMTR) 522 receives and processes (e.g., OFDM modulates and conditions) a respective transmit symbol stream to generate a downlink signal. N_{ap} transmitter units 522a through 522ap provide N_{ap} downlink signals for transmission from N_{ap} antennas 524a through 524ap, respectively.

[0063] At user terminal 120, N_{ut} antennas 552a through 552ut receive the N_{ap} downlink signals, and each antenna provides a received signal to a respective receiver unit (RCVR) 554. Each receiver unit 554 performs processing (e.g., conditioning and OFDM demodulation) complementary to that performed by transmitter units 522 and provides a stream of received symbols. A receive (RX) spatial processor 560 performs spatial processing on N_{ut} received symbol streams from N_{ut} receiver units 554 and provides N_S streams of recovered data symbols. The processing by RX spatial processor 560 is dependent on whether the steered or non-steered mode is selected for use and may be performed as described above. An RX data processor 570 processes (e.g., demaps, deinterleaves, and decodes) the N_S recovered data symbol streams to obtain N_S decoded data streams, which may be provided to a data sink 572 for storage and/or a controller 580 for further processing.

[0064] A channel estimator 578 estimates the downlink channel response based on received pilot symbols and provides channel estimates, which may include channel gain estimates, SNR estimates, and so on. Controller 580 receives the channel estimates, derives the matrices used by RX spatial processor 560 and a TX spatial processor 590 for spatial processing, and determines a suitable rate for each data symbol stream sent on the downlink. The rates and uplink data are processed by a TX data processor 588, spatially processed (as required) by TX spatial processor 590, multiplexed with pilot symbols, conditioned by N_{ut} transmitter units 554a through 554ut, and transmitted via antennas 552a through 552ut.

[0065] At access point 110, the N_{ut} transmitted uplink signals are received by antennas 524, conditioned and demodulated by receiver units 522, and processed by an RX spatial processor 540 and an RX data processor 542. The rates are provided to controller 530 and used to control data transmission on the downlink.

[0066] Access point 110 and user terminal 120 may perform similar or different processing for uplink data and pilot transmission.

[0067] Controllers 530 and 580 control the operation of various processing units at access point 110 and user terminal 120, respectively. SM mode selectors 534 and 584 select the appropriate spatial multiplexing mode to use for access point 110 and user terminal 120, respectively, based on various factors such as those described above. Memory units 532 and 582 store data and program codes used by controllers 530 and 580, respectively.

[0068] **FIG. 6** shows a flow diagram of a process 600 for transmitting and receiving data in the MIMO system. Process 600 may be performed by a user terminal and an access point for data transmission on the downlink and uplink.

[0069] Initially, an SM mode is selected from among multiple supported SM modes, which may include the steered and non-steered modes described above (step 612). The mode selection may be based on the calibration status of the terminal, the amount of data to send, the SNR and/or channel conditions, the capability of the other communicating entity, and so on. The selected SM mode may also change during a data session.

[0070] For data transmission (block 620), multiple data streams for a first communication link (e.g., the uplink) are coded and modulated in accordance with their selected rates to obtain multiple data symbol streams for the first link (step 622). These data symbol streams are then spatially processed in accordance with the selected SM mode to obtain multiple transmit symbol streams for transmission from multiple antennas and via the first link (step 624). The transmit spatial processing is with a matrix of steering vectors for the steered mode and with the identity matrix for the non-steered mode.

[0071] For data reception (block 630), multiple received symbol streams, obtained from the multiple antennas for a second communication link (e.g., the downlink), are spatially processed in accordance with the selected SM mode to obtain multiple recovered data symbol streams (step 632). The receive spatial processing is with a matrix of

eigenvectors for the steered mode and a spatial filter matrix for the non-steered mode. The spatial filter matrix may be derived based on the CCMI, MMSE, SIC, or some other technique. The recovered data symbol streams are then demodulated and decoded in accordance with their selected rates to obtain multiple decoded data streams for the second link (step 634).

[0072] The data transmission in block 620 and the data reception in block 630 may occur simultaneously or at different times. Pilots and rates are also transmitted and received in order to support data transmission and reception with the selected SM mode.

[0073] The multi-mode terminal and access point and the data transmission/reception techniques described herein may be implemented by various means. For example, these entities and techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the processing units for these entities and techniques may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[0074] For a software implementation, the techniques described herein may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory units 532 and 582 in FIG. 5) and executed by a processor (e.g., controllers 530 and 580). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[0075] Headings are included herein for reference and to aid in locating certain sections. These headings are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[0076] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be

limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

[0077] **WHAT IS CLAIMED IS:**

CLAIMS

1. A terminal in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

a mode selector operable to select a spatial multiplexing mode from among a plurality of spatial multiplexing modes supported by the terminal, wherein each of the plurality of spatial multiplexing modes supports simultaneous transmission of multiple data symbol streams via multiple spatial channels of a MIMO channel formed with a plurality of antennas at the terminal;

a transmit spatial processor operable to spatially process a first plurality of data symbol streams in accordance with the selected spatial multiplexing mode to obtain a plurality of transmit symbol streams for transmission from the plurality of antennas and via a first communication link; and

a receive spatial processor operable to spatially process a plurality of received symbol streams, obtained from the plurality of antennas, in accordance with the selected spatial multiplexing mode to obtain a plurality of recovered data symbol streams, which are estimates of a second plurality of data symbol streams sent via a second communication link.

2. The terminal of claim 1, wherein the plurality of spatial multiplexing modes include a steered mode and a non-steered mode.

3. The terminal of claim 2, wherein the steered mode supports simultaneous transmission of multiple data symbol streams via multiple orthogonal spatial channels of the MIMO channel, and wherein the non-steered mode supports simultaneous transmission of multiple data symbol streams from the plurality of antennas.

4. The terminal of claim 2, wherein

the transmit spatial processor is operable to multiply the first plurality of data symbol streams with a matrix of steering vectors for the steered mode and with an identity matrix for the non-steered mode, and

the receive spatial processor is operable to multiply the plurality of received symbol streams with a matrix of eigenvectors for the steered mode and with a spatial filter matrix for the non-steered mode.

5. The terminal of claim 4, further comprising:

a channel estimator operable to estimate a channel response of the second communication link; and

a controller operable to derive the spatial filter matrix based on the estimated channel response for the second communication link.

6. The terminal of claim 5, wherein the controller is operable to derive the spatial filter matrix based on a channel correlation matrix inversion (CCMI) technique or a minimum mean square error (MMSE) technique.

7. The terminal of claim 5, wherein the controller is operable to derive the spatial filter matrix based on a successive interference cancellation (SIC) technique and using a channel correlation matrix inversion (CCMI) technique or a minimum mean square error (MMSE) technique.

8. The terminal of claim 2, further comprising:

a transmit data processor operable to code and modulate a first plurality of data streams in accordance with a first plurality of rates to obtain the first plurality of data symbol streams for the first communication link; and

a receive data processor operable to demodulate and decode the plurality of recovered data symbol streams in accordance with a second plurality of rates to obtain a plurality of decoded data streams for the second communication link.

9. The terminal of claim 8, wherein the first plurality of rates are for a plurality of eigenmodes of the MIMO channel for the steered mode and are for the plurality of antennas for the non-steered mode.

10. The terminal of claim 2, wherein the mode selector is operable to select the steered mode if the terminal is calibrated and the non-steered mode if the terminal is not calibrated, and wherein channel response of the second communication link is reciprocal of channel response of the first communication link if the terminal is calibrated.

11. The terminal of claim 2, wherein the mode selector is operable to select the steered mode or the non-steered mode based on an amount of data to send, channel conditions, capability of an entity in communication with the terminal, or a combination thereof.

12. The terminal of claim 2, wherein the mode selector is operable to select the non-steered mode for a first portion of a data session and to select the steered mode for a remaining portion of the data session.

13. The terminal of claim 2, wherein the mode selector is operable to select the steered mode or the non-steered mode based on received signal-to-noise-and-interference ratio (SNR).

14. The terminal of claim 2, wherein the transmit spatial processor is further operable to multiplex a steered pilot for the steered mode and an unsteered pilot for the non-steered mode, wherein the steered pilot is transmitted on eigenmodes of the MIMO channel, and wherein the unsteered pilot comprises a plurality of orthogonal pilot transmissions from the plurality of antennas.

15. The terminal of claim 2, wherein the transmit spatial processor is further operable to multiplex an unsteered pilot for both the steered and non-steered modes, and wherein the unsteered pilot comprises a plurality of orthogonal pilot transmissions from the plurality of antennas.

16. The terminal of claim 1 and operable to communicate with an access point in the MIMO system.

17. The terminal of claim 1 and operable to communicate peer-to-peer with another terminal in the MIMO system.

18. The terminal of claim 1, wherein the MIMO system utilizes orthogonal frequency division multiplexing (OFDM), and wherein the transmit and receive spatial processors are operable to perform spatial processing for each of a plurality of subbands.

19. The terminal of claim 1, wherein the MIMO system is a time division duplex (TDD) system.

20. A method of processing data in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

selecting a spatial multiplexing mode from among a plurality of spatial multiplexing modes, wherein each of the plurality of spatial multiplexing modes supports simultaneous transmission of multiple data symbol streams via multiple spatial channels of a MIMO channel;

spatially processing a first plurality of data symbol streams in accordance with the selected spatial multiplexing mode to obtain a plurality of transmit symbol streams for transmission from a plurality of antennas and via a first communication link; and

spatially processing a plurality of received symbol streams, obtained from the plurality of antennas, in accordance with the selected spatial multiplexing mode to obtain a plurality of recovered data symbol streams, which are estimates of a second plurality of data symbol streams sent via a second communication link.

21. The method of claim 20, wherein the plurality of spatial multiplexing modes include a steered mode and a non-steered mode, the steered mode supporting simultaneous transmission of multiple data symbol streams via multiple orthogonal spatial channels of the MIMO channel, and the non-steered mode supporting simultaneous transmission of multiple data symbol streams from the plurality of antennas.

22. The method of claim 21, wherein the first plurality of data symbol streams are multiplied with a matrix of steering vectors for the steered mode and with an identity matrix for the non-steered mode, and wherein the plurality of received symbol streams are multiplied with a matrix of eigenvectors for the steered mode and with a spatial filter matrix for the non-steered mode.

23. The method of claim 22, further comprising:

estimating a channel response of the second communication link; and

deriving the spatial filter matrix based on the estimated channel response for the second communication link.

24. The method of claim 23, wherein the spatial filter matrix is derived based on a channel correlation matrix inversion (CCMI) technique, a minimum mean square error (MMSE) technique, or a successive interference cancellation (SIC) technique.

25. An apparatus in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

means for selecting a spatial multiplexing mode from among a plurality of spatial multiplexing modes, wherein each of the plurality of spatial multiplexing modes supports simultaneous transmission of multiple data symbol streams via multiple spatial channels of a MIMO channel;

means for spatially processing a first plurality of data symbol streams in accordance with the selected spatial multiplexing mode to obtain a plurality of transmit symbol streams;

means for transmitting the plurality of transmit symbol streams from a plurality of antennas and via a first communication link;

means for receiving a plurality of received symbol streams from the plurality of antennas for a second communication link; and

means for spatially processing the plurality of received symbol streams in accordance with the selected spatial multiplexing mode to obtain a plurality of recovered data symbol streams, which are estimates of a second plurality of data symbol streams sent via the second communication link.

26. The apparatus of claim 25, wherein the plurality of spatial multiplexing modes include a steered mode and a non-steered mode, the steered mode supporting simultaneous transmission of multiple data symbol streams via multiple orthogonal spatial channels of the MIMO channel, and the non-steered mode supporting simultaneous transmission of multiple data symbol streams from the plurality of antennas.

27. The apparatus of claim 26, wherein the first plurality of data symbol streams are multiplied with a matrix of steering vectors for the steered mode and with an identity matrix for the non-steered mode, and wherein the plurality of received symbol streams are multiplied with a matrix of eigenvectors for the steered mode and with a spatial filter matrix for the non-steered mode.

28. The apparatus of claim 27, further comprising:

means for estimating a channel response of the second communication link; and

means for deriving the spatial filter matrix based on the estimated channel response for the second communication link.

29. The apparatus of claim 28, wherein the spatial filter matrix is derived based on a channel correlation matrix inversion (CCMI) technique, a minimum mean square error (MMSE) technique, or a successive interference cancellation (SIC) technique.

30. An access point in a wireless multiple-input multiple-output (MIMO) communication system, comprising:

a mode selector operable to select a spatial multiplexing mode from among a plurality of spatial multiplexing modes supported by the access point, wherein each of the plurality of spatial multiplexing modes supports simultaneous transmission of multiple data symbol streams via multiple spatial channels of a MIMO channel formed with a plurality of antennas at the access point;

a transmit spatial processor operable to spatially process a first plurality of data symbol streams in accordance with the selected spatial multiplexing mode to obtain a plurality of transmit symbol streams for transmission from the plurality of antennas and via a first communication link; and

a receive spatial processor operable to spatially process a plurality of received symbol streams, obtained from the plurality of antennas, in accordance with the selected spatial multiplexing mode to obtain a plurality of recovered data symbol streams, which are estimates of a second plurality of data symbol streams sent via a second communication link.

31. The access point of claim 30, wherein the plurality of spatial multiplexing modes include a steered mode and a non-steered mode.

32. The access point of claim 31, wherein

the transmit spatial processor is operable to multiply the first plurality of data symbol streams with a matrix of steering vectors for the steered mode and with an identity matrix for the non-steered mode, and

the receive spatial processor is operable to multiply the plurality of received symbol streams with a matrix of eigenvectors for the steered mode and with a spatial filter matrix for the non-steered mode.

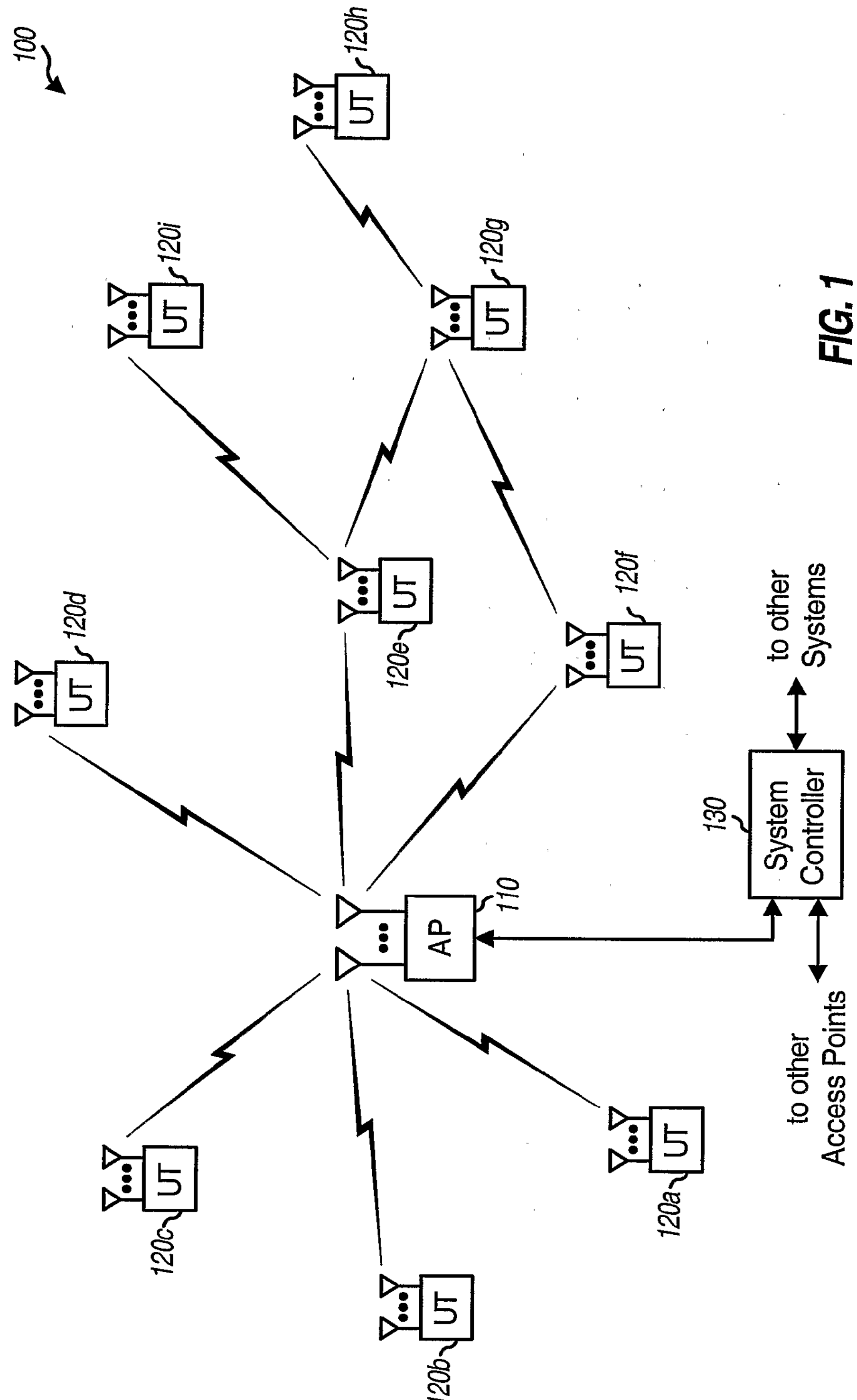
100
UT
120h
UT
120i
UT
120g
UT
120d
UT
120e
UT
120f
to other
Systems
130
System
Controller
AP
110
to other
Access Points
UT
120c
UT
120a
UT
120b

FIG. 1

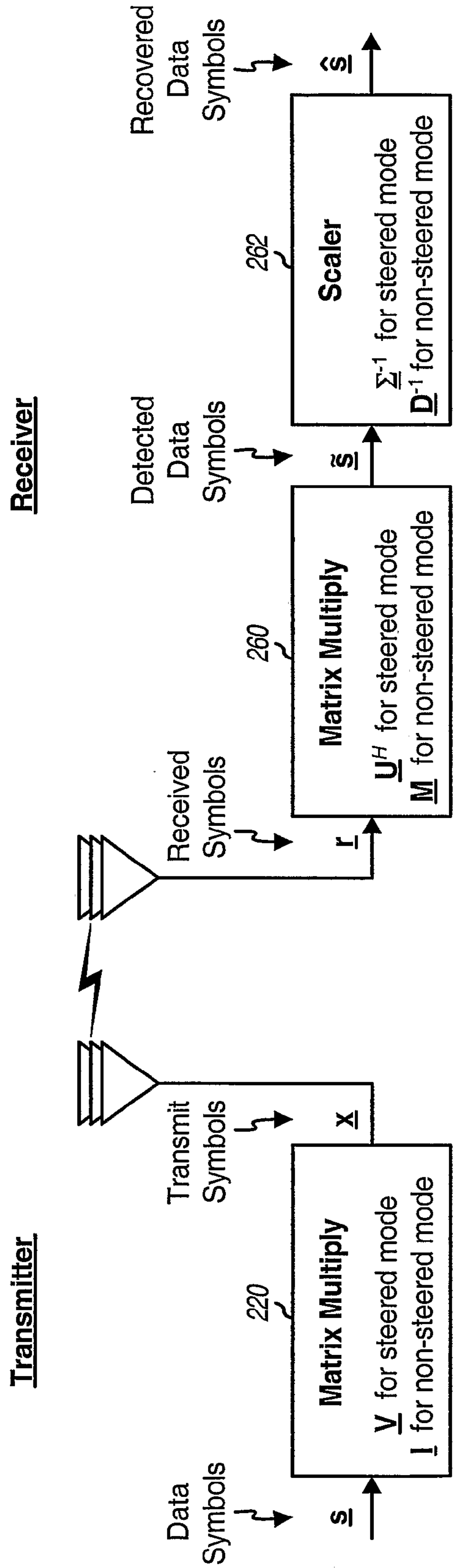
Transmitter
Receiver
Data Symbols
s
220
Matrix Multiply
V for steered mode
I for non-steered mode
Transmit Symbols
x
Received Symbols
r
260
Matrix Multiply
$\underline{U}^H$ for steered mode
M for non-steered mode
Detected Data Symbols
$\tilde{\underline{s}}$
262
Scaler
$\underline{\Sigma}^{-1}$ for steered mode
$\underline{D}^{-1}$ for non-steered mode
Recovered Data Symbols
$\hat{\underline{s}}$

FIG. 2

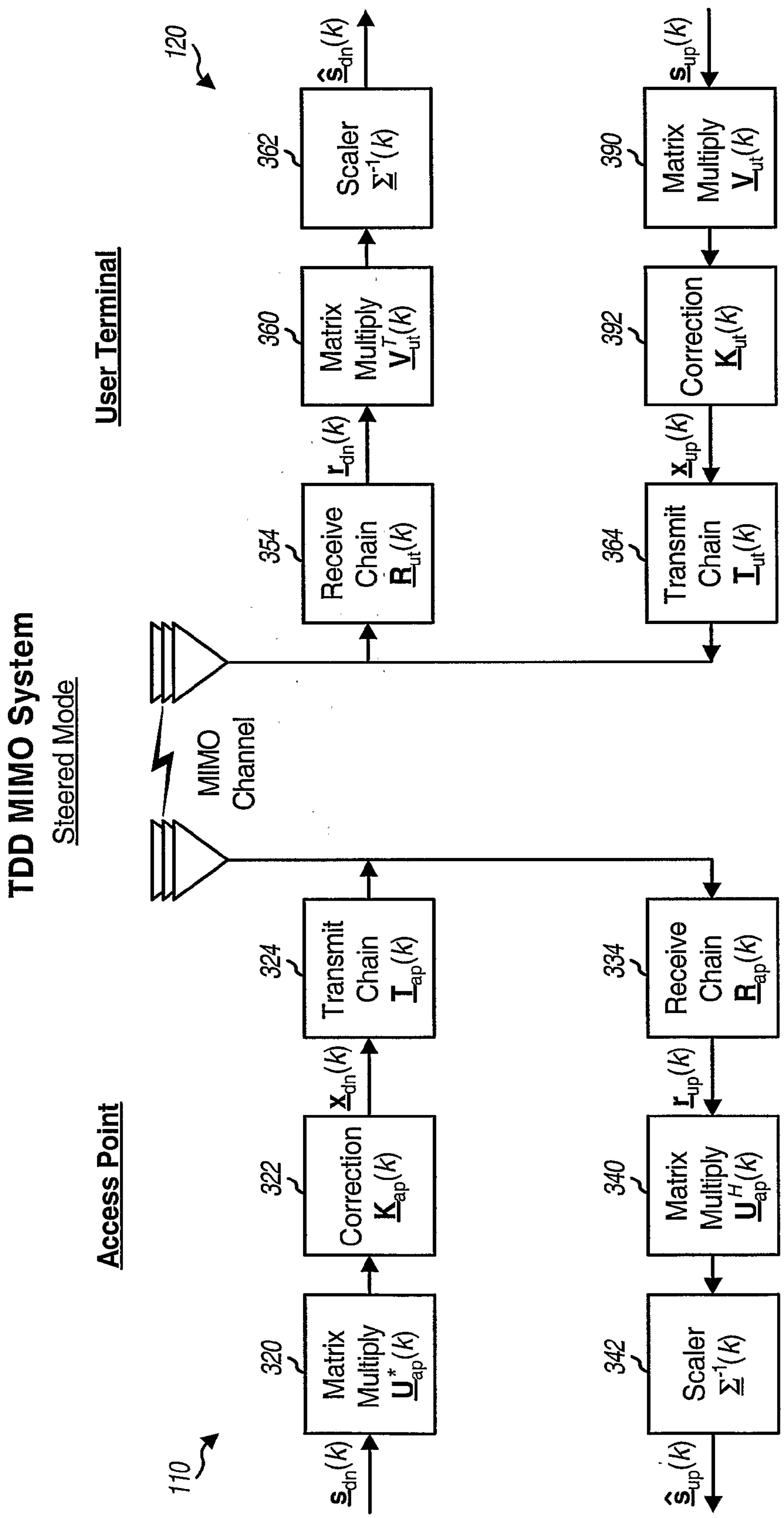
TDD MIMO System
Steered Mode
Access Point
User Terminal
110
120
MIMO Channel
320 Matrix Multiply $\underline{\mathbf{U}}^{*}_{ap}(k)$
322 Correction $\underline{\mathbf{K}}_{ap}(k)$
324 Transmit Chain $\underline{\mathbf{T}}_{ap}(k)$
334 Receive Chain $\underline{\mathbf{R}}_{ap}(k)$
340 Matrix Multiply $\underline{\mathbf{U}}^{H}_{ap}(k)$
342 Scaler $\underline{\Sigma}^{-1}(k)$
354 Receive Chain $\underline{\mathbf{R}}_{ut}(k)$
360 Matrix Multiply $\underline{\mathbf{V}}^{T}_{ut}(k)$
362 Scaler $\underline{\Sigma}^{-1}(k)$
364 Transmit Chain $\underline{\mathbf{T}}_{ut}(k)$
392 Correction $\underline{\mathbf{K}}_{ut}(k)$
390 Matrix Multiply $\underline{\mathbf{V}}_{ut}(k)$
$\underline{\mathbf{s}}_{dn}(k)$
$\underline{\mathbf{x}}_{dn}(k)$
$\underline{\mathbf{r}}_{dn}(k)$
$\hat{\underline{\mathbf{s}}}_{dn}(k)$
$\underline{\mathbf{s}}_{up}(k)$
$\underline{\mathbf{x}}_{up}(k)$
$\underline{\mathbf{r}}_{up}(k)$
$\hat{\underline{\mathbf{s}}}_{up}(k)$

FIG. 3

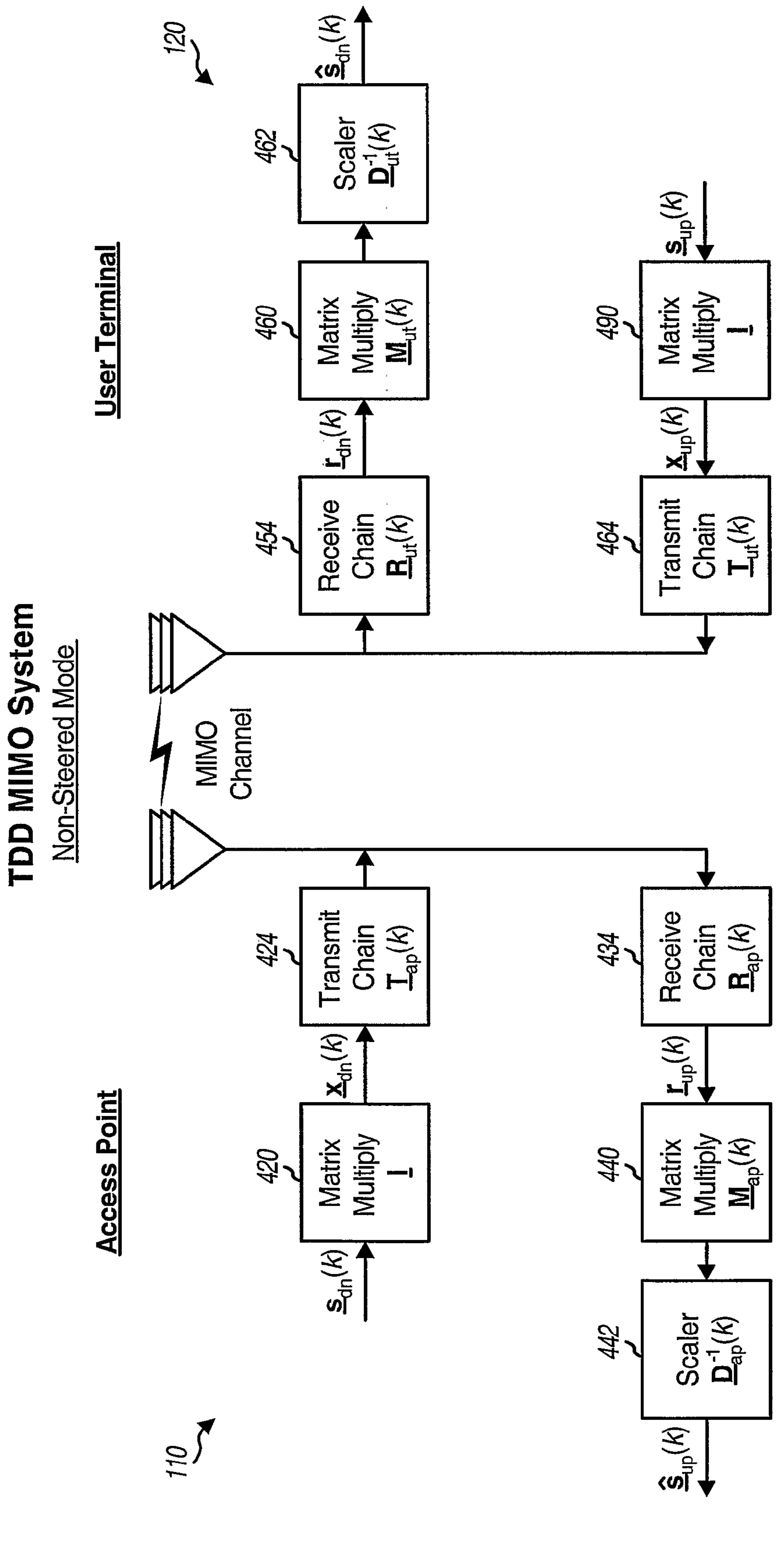
TDD MIMO System
Non-Steered Mode
Access Point
User Terminal
110
120
MIMO Channel
420 Matrix Multiply I
424 Transmit Chain T_ap(k)
434 Receive Chain R_ap(k)
440 Matrix Multiply M_ap(k)
442 Scaler D_ap^-1(k)
454 Receive Chain R_ut(k)
460 Matrix Multiply M_ut(k)
462 Scaler D_ut^-1(k)
464 Transmit Chain T_ut(k)
490 Matrix Multiply I
s_dn(k)
x_dn(k)
r_up(k)
ŝ_up(k)
r_dn(k)
ŝ_dn(k)
s_up(k)
x_up(k)

FIG. 4

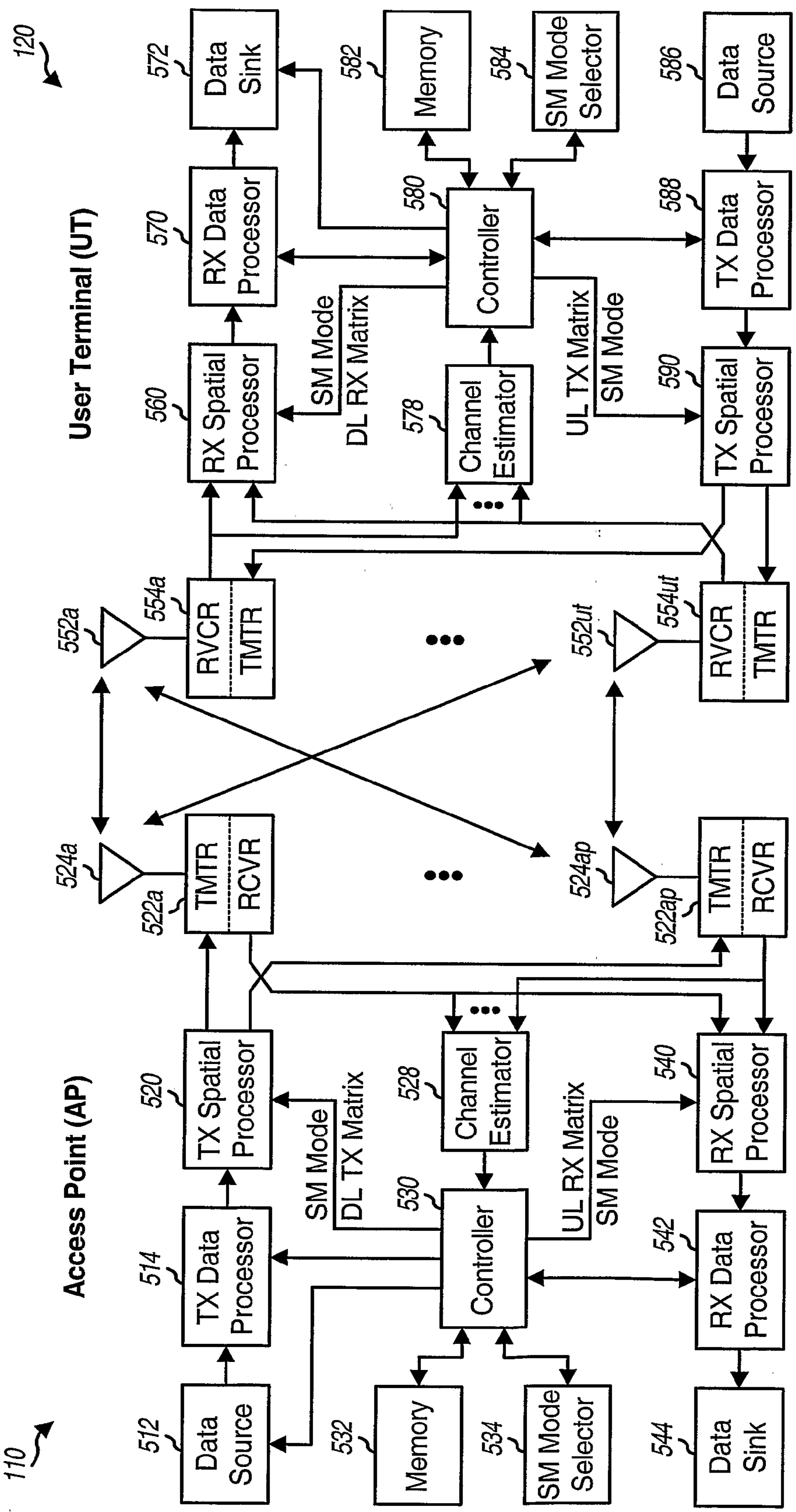
110
Access Point (AP)
512
Data Source
514
TX Data Processor
520
TX Spatial Processor
SM Mode
DL TX Matrix
522a
TMTR
RCVR
524a
522ap
524ap
528
Channel Estimator
530
Controller
532
Memory
534
SM Mode Selector
UL RX Matrix
SM Mode
540
RX Spatial Processor
542
RX Data Processor
544
Data Sink
120
User Terminal (UT)
552a
554a
RVCR
TMTR
552ut
554ut
560
RX Spatial Processor
SM Mode
DL RX Matrix
570
RX Data Processor
572
Data Sink
578
Channel Estimator
580
Controller
582
Memory
584
SM Mode Selector
UL TX Matrix
SM Mode
586
Data Source
588
TX Data Processor
590
TX Spatial Processor

FIG. 5

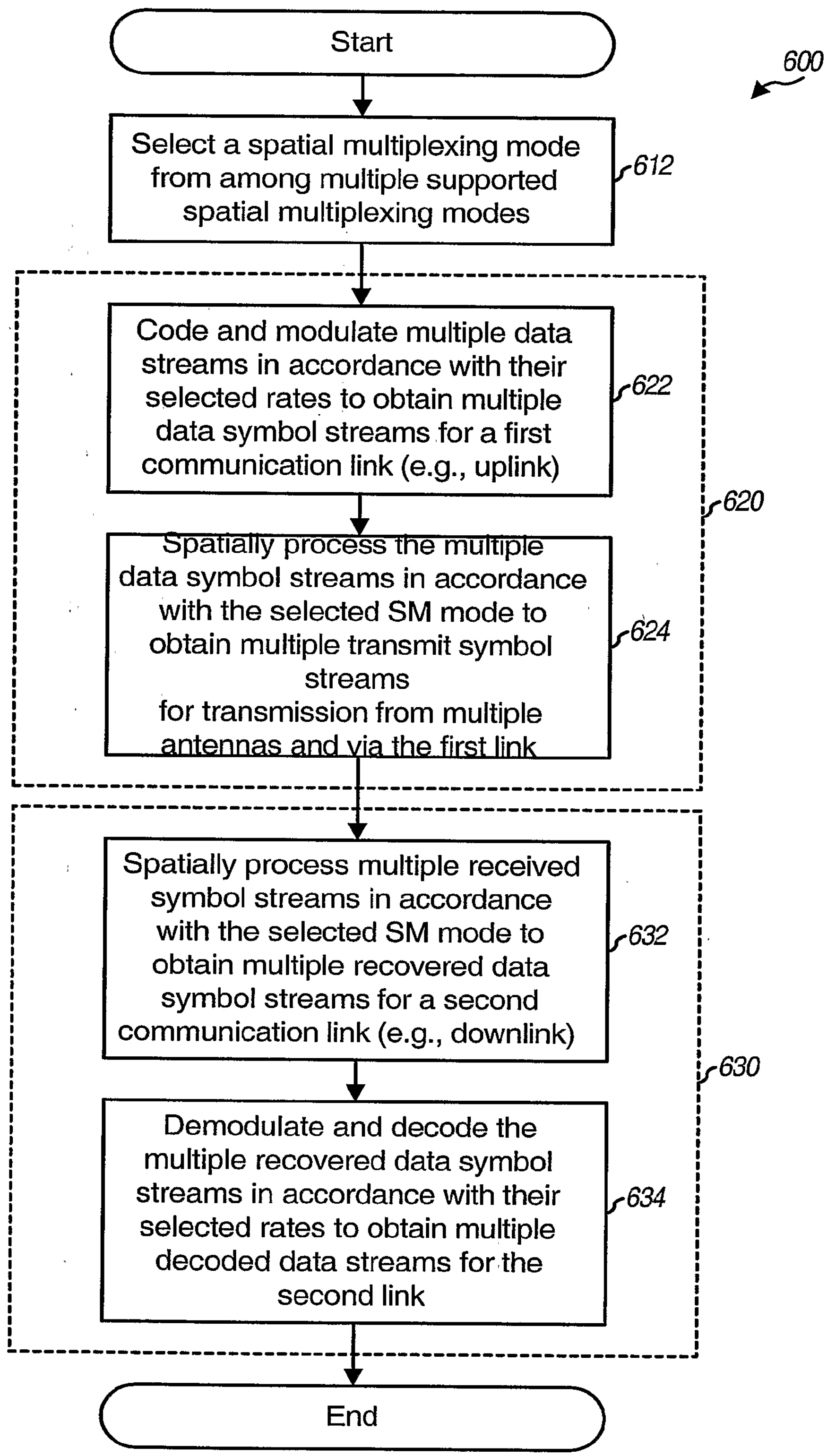
Start
600
Select a spatial multiplexing mode from among multiple supported spatial multiplexing modes
612
Code and modulate multiple data streams in accordance with their selected rates to obtain multiple data symbol streams for a first communication link (e.g., uplink)
622
Spatially process the multiple data symbol streams in accordance with the selected SM mode to obtain multiple transmit symbol streams for transmission from multiple antennas and via the first link
624
620
Spatially process multiple received symbol streams in accordance with the selected SM mode to obtain multiple recovered data symbol streams for a second communication link (e.g., downlink)
632
Demodulate and decode the multiple recovered data symbol streams in accordance with their selected rates to obtain multiple decoded data streams for the second link
634
630
End

FIG. 6

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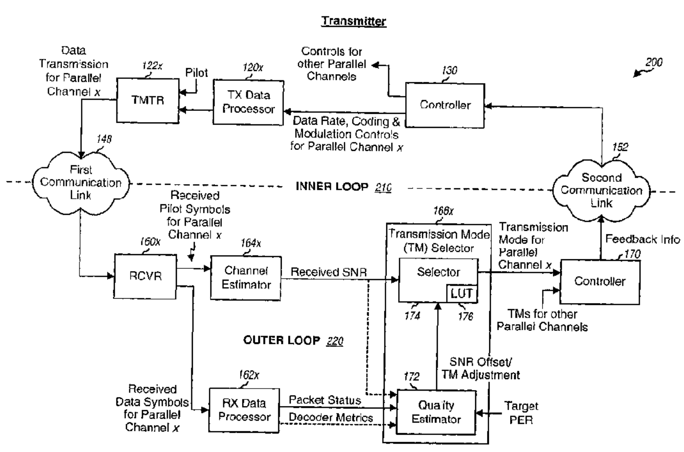
(54) Title: CLOSED-LOOP RATE CONTROL FOR A MULTI-CHANNEL COMMUNICATION SYSTEM



(57) Abstract: Closed-loop rate control for data transmission on multiple parallel channels is provided. An inner loop estimates the channel conditions for a communication link and selects a suitable data rate for each of the multiple parallel channels based on the channel estimates. For each parallel channel, a received SNR is computed based on the channel estimates, an operating SNR is computed based on the received SNR and an SNR offset for the parallel channel, and the data rate is selected based on the operating SNR for the parallel channel and a set of required SNRs for a set of data rates supported by the system. An outer loop estimates the quality of data transmissions received on the multiple parallel channels and adjusts the operation of the inner loop. For example, the SNR offset for each parallel channel is adjusted based on the status of packets received on that parallel channel.

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CLOSED-LOOP RATE CONTROL FOR A MULTI-CHANNEL COMMUNICATION SYSTEM

BACKGROUND

I. Field

[1001] The present invention relates generally to data communication, and more specifically to techniques for performing rate control for data transmission on multiple parallel channels in a multi-channel communication system.

II. Background

[1002] A multi-channel communication system utilizes multiple "parallel channels" for data transmission. These parallel channels may be formed in the time domain, frequency domain, spatial domain, or a combination thereof. For example, the multiple parallel channels may be formed by different time slots in a time division multiplex (TDM) communication system, different frequency subbands in a frequency division multiplex (FDM) communication system, different disjoint sets of subbands in an orthogonal frequency division multiplex (OFDM) communication system, or different spatial channels in a multiple-input multiple-output (MIMO) communication system. TDM, FDM, OFDM, and MIMO systems are described in further detail below.

[1003] The multiple parallel channels may experience different channel conditions (e.g., different fading, multipath, and interference effects) and may achieve different signal-to-noise ratios (SNRs). The SNR of a parallel channel determines its transmission capability, which is typically quantified by a particular data rate that may be reliably transmitted on the parallel channel. If the SNR varies from parallel channel to parallel channel, then the supported data rate would also vary from channel to channel. Moreover, since the channel conditions typically vary with time, the data rates supported by the multiple parallel channels also vary with time.

[1004] Rate control is a major challenge in a multi-channel communication system that experiences continually varying channel conditions. Rate control entails controlling the data rate of each of the multiple parallel channels based on the channel conditions. The goal of the rate control should be to maximize the overall throughput

on the multiple parallel channels while meeting certain quality objectives, which may be quantified by a particular packet error rate (PER) or some other criterion.

[1005] There is therefore a need in the art for techniques to effectively perform rate control for multiple parallel channels having varying SNRs.

SUMMARY

[1006] Techniques for performing closed-loop rate control for data transmission on multiple parallel channels are described herein. Closed-loop rate control may be achieved with one or multiple loops. An inner loop estimates the channel conditions for a communication link and selects a suitable data rate for each of the multiple parallel channels (e.g., to achieve high overall throughput). An outer loop (which is optional) estimates the quality of the data transmissions received on the multiple parallel channels and adjusts the operation of the inner loop.

[1007] For the inner loop, channel estimates are initially obtained for the multiple parallel channels (e.g., based on received pilot symbols). The channel estimates may include channel gain estimates for multiple subbands of each parallel channel, an estimate of the noise floor at the receiver, and so on. A suitable "transmission mode" is then selected for each parallel channel based on (1) the transmit power allocated to the parallel channel, (2) the channel estimates for the parallel channel, (3) an SNR offset provided by the outer loop for the parallel channel, and (4) other information provided by the outer loop. A transmission mode indicates, among other things, a specific data rate to use for a parallel channel. The SNR offset indicates the amount of back-off to use for the parallel channel and influences the selection of the transmission mode for the parallel channel. The other information from the outer loop may direct the inner loop to select a transmission mode with a data rate lower than that normally selected for the parallel channel, for example, if excessive packet errors are received for the parallel channel. The transmitter and receiver process data for each parallel channel in accordance with the transmission mode selected for that parallel channel.

[1008] For the outer loop, the receiver estimates the quality of the data transmissions received via the multiple parallel channels. For example, the receiver may determine the status of each received data packet (e.g., as good or bad, as described below), obtain decoder metrics for each data stream, estimate the received SNR for each parallel channel, and so on. The outer loop then adjusts the operation of the inner loop

for each parallel channel based on the estimated received quality for that parallel channel. For example, the outer loop may adjust the SNR offset for each parallel channel to achieve a target packet error rate (PER) for that parallel channel. The outer loop may also direct the inner loop to select a transmission mode with a lower data rate for a parallel channel if excessive packet errors are detected for that parallel channel.

[1009] Various aspects and embodiments of the invention are also described in further detail below.

BRIEF DESCRIPTION OF THE DRAWINGS

[1010] The features, nature, and advantages of the present invention will become more apparent from the detailed description set forth below when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

[1011] FIG. 1 shows a transmitter and a receiver in a multi-channel communication system with closed-loop rate control for N_C parallel channels;

[1012] FIG. 2 shows a closed-loop rate control mechanism;

[1013] FIG. 3 shows an exemplary process to transmit N_C data streams on N_C parallel channels using N_C transmission modes selected with closed-loop rate control;

[1014] FIG. 4 shows an exemplary process for the outer loop;

[1015] FIG. 5 shows an exemplary TDD MIMO-OFDM system;

[1016] FIG. 6 shows a frame structure used in the TDD MIMO-OFDM system;

[1017] FIG. 7 shows a process for transmitting multiple data streams on multiple wideband eigenmodes on the downlink and uplink in the TDD MIMO-OFDM system;

[1018] FIG. 8 shows a process for selecting N_S transmission modes for N_S wideband eigenmodes;

[1019] FIGS. 9A and 9B show an access point and a terminal in the TDD MIMO-OFDM system for downlink and uplink transmission, respectively;

[1020] FIG. 10 shows a transmitter subsystem;

[1021] FIG. 11 shows a receiver subsystem; and

[1022] FIGS. 12A and 12B show exemplary timing diagrams for closed-loop rate control for the downlink and uplink, respectively.

DETAILED DESCRIPTION

[1023] The word "exemplary" is used herein to mean "serving as an example, instance, or illustration." Any embodiment or design described herein as "exemplary" is not necessarily to be construed as preferred or advantageous over other embodiments or designs.

[1024] As used herein, "rate control" entails controlling the data rate of each of multiple parallel channels based on channel conditions. The data rate for each parallel channel is determined by the transmission mode selected for use for that parallel channel. Rate control may thus be achieved by controlling the transmission modes used for the multiple parallel channels.

[1025] **FIG. 1** shows a block diagram of a transmitter 110 and a receiver 150 in a multi-channel communication system 100 with closed-loop rate control for N_C parallel channels, where $N_C > 1$. The N_C parallel channels may be formed in various manners, as described below. For downlink transmission, transmitter 110 is an access point, receiver 150 is a user terminal, first communication link 148 is the downlink (i.e., forward link), and second communication link 152 is the uplink (i.e., reverse link). For uplink transmission, transmitter 110 is a user terminal, receiver 150 is an access point, and the first and second communication links are the uplink and downlink, respectively.

[1026] At transmitter 110, a transmit (TX) data processor 120 receives N_C data streams, one stream for each of the N_C parallel channels. Each parallel channel is associated with a specific transmission mode that indicates a set of transmission parameters to use for that parallel channel. A transmission mode may indicate (or may be associated with) a particular data rate, a particular coding scheme or code rate, a particular interleaving scheme, a particular modulation scheme, and so on, to use for data transmission. An exemplary set of transmission modes is given in Table 2 below. For each parallel channel, the data rate is indicated by a data rate control, the coding scheme is indicated by a coding control, and the modulation scheme is indicated by a modulation control. These controls are provided by a controller 130 and are generated based on the transmission mode selected for each parallel channel using feedback information obtained from receiver 150 and possibly other information (e.g., channel estimates) obtained by transmitter 110.

[1027] TX data processor 120 codes, interleaves, and modulates each data stream in accordance with the transmission mode selected for its parallel channel to provide a

corresponding stream of modulation symbols. TX data processor 120 provides N_C modulation symbol streams for the N_C data streams. A transmitter unit (TMTR) 122 then processes the N_C modulation symbol streams in a manner specified by the system. For example, transmitter unit 122 may perform OFDM processing for an OFDM system, spatial processing for a MIMO system, or both spatial and OFDM processing for a MIMO-OFDM system (which is a MIMO system that utilizes OFDM). A pilot is also transmitted to assist receiver 150 in performing a number of functions such as channel estimation, acquisition, frequency and timing synchronization, coherent demodulation, and so on. Transmitter unit 122 multiplexes pilot symbols with the modulation symbols for each parallel channel, processes the multiplexed symbols, and provides a modulated signal for each antenna used for data transmission. Each modulated signal is then transmitted via first communication link 148 to receiver 150. First communication link 148 distorts each modulated signal with a particular channel response and further degrades the modulated signal with (1) additive white Gaussian noise (AWGN) having a variance of N_0 and (2) possibly interference from other transmitters.

[1028] At receiver 150, the transmitted signal(s) are received by one or more receive antennas, and the received signal from each antenna is provided to a receiver unit (RCVR) 160. Receiver unit 160 conditions and digitizes each received signal to provide a corresponding stream of samples. Receiver unit 160 further processes the samples in a manner that is complementary to that performed by transmitter unit 122 to provide N_C streams of "recovered" symbols, which are estimates of the N_C streams of modulation symbols sent by transmitter 110.

[1029] A receive (RX) data processor 162 then processes the N_C recovered symbol streams in accordance with the N_C transmission modes selected for the N_C parallel channels to obtain N_C decoded data streams, which are estimates of the N_C data streams sent by transmitter 110. The processing by RX data processor 162 may include demodulation, deinterleaving, and decoding. RX data processor 162 may further provide the status of each received data packet and/or decoder metrics for each decoded data stream.

[1030] Receiver unit 160 also provides received pilot symbols for the N_C parallel channels to a channel estimator 164. Channel estimator 164 processes these received pilot symbols to obtain channel estimates for the N_C parallel channels. The channel

estimates may include, for example, channel gain estimates, noise variance N_0 estimate, and so on. The noise variance N_0, which is the noise floor observed at receiver 150, includes channel noise, receiver circuitry noise, interference (i.e., cross-talk) from other transmitting entities, and so on.

[1031] A transmission mode (TM) selector 166 receives the channel estimates from channel estimator 164 and possibly packet status and/or decoder metrics from RX data processor 162. Transmission mode selector 166 computes an operating SNR for each of the N_C parallel channels based on the channel estimates and an SNR offset for that parallel channel. Transmission mode selector 166 then selects a suitable transmission mode for each parallel channel based on the operating SNR and outer loop information for the parallel channel. The transmission mode selection is described in detail below.

[1032] A controller 170 receives the N_C selected transmission modes, TM 1 through TM N_C, from transmission mode selector 166 and the packet status from RX data processor 162 (not shown). Controller 170 then assembles feedback information for transmitter 110. The feedback information may include the N_C selected transmission modes for the N_C parallel channels, acknowledgments (ACKs) and/or negative acknowledgments (NAKs) for received data packets, a pilot, and/or other information. The feedback information is then sent via second communication link 152 to transmitter 110. Transmitter 110 uses the feedback information to adjust the processing of the N_C data streams sent to receiver 150. For example, transmitter 110 may adjust the data rate, the coding scheme, the modulation scheme, or any combination thereof, for each of the N_C data streams sent on the N_C parallel channels to receiver 150. The feedback information is used to increase the efficiency of the system by allowing data to be transmitted at the best-known settings supported by first communication link 148.

[1033] In the embodiment shown in FIG. 1, the channel estimation and transmission mode selection are performed by receiver 150 and the N_C transmission modes selected for the N_C parallel channels are sent back to transmitter 110. In other embodiments, the channel estimation and transmission mode selection may be performed (1) by transmitter 110 based on feedback information obtained from receiver 150 and/or other information obtained by transmitter 110 or (2) jointly by both transmitter 110 and receiver 150.

[1034] **FIG. 2** shows a block diagram of an embodiment of a closed-loop rate control mechanism 200, which includes an inner loop 210 that operates in conjunction

with an outer loop 220. For simplicity, the operation of inner loop 210 and outer loop 220 for only one parallel channel x is shown in FIG. 2. In general, the same processing may be performed independently for each of the N_C parallel channels.

[1035] For inner loop 210, channel estimator 164x estimates the channel conditions for parallel channel x and provides channel estimates (e.g., channel gain estimates and noise floor estimate). A selector 174 within transmission mode selector 166x computes a received SNR for parallel channel x based on (1) the channel estimates from channel estimator 164x and (2) an SNR offset and/or a transmission mode adjustment for parallel channel x from a quality estimator 172. For clarity, the received SNR is symbolically shown as being provided by channel estimator 164x to selector 174 in FIG. 2. Selector 174 then selects a transmission mode for parallel channel x based on the received information, as described below. The select transmission mode for parallel channel x is included in the feedback information sent by controller 170 to the transmitter. At the transmitter, controller 130 receives the selected transmission mode for parallel channel x and determines the data rate, coding, and modulation controls for parallel channel x. Data is then processed in accordance with these controls by TX data processor 120x, further multiplexed with pilot symbols and conditioned by transmitter unit 122x, and sent to the receiver. The channel estimation and transmission mode selection may be performed periodically, at scheduled times, whenever changes in the communication link are detected, only as necessary (e.g., prior to and during data transmission), or at other times.

[1036] Outer loop 220 estimates quality of the data transmission received on parallel channel x and adjusts the operation of inner loop 210 for parallel channel x. The received data symbols for parallel channel x are processed by RX data processor 162x, and the status of each received packet on parallel channel x and/or decoder metrics are provided to quality estimator 172. The decoder metrics may include a re-encoded symbol error rate (SER), a re-encoded power metric, a modified Yamamoto metric (for a convolutional decoder), minimum or average log-likelihood ratio (LLR) among bits in a decoded packet (for a Turbo decoder), and so on. The re-encoded SER is the error rate between the received symbols from receiver unit 160 and the re-encoded symbols obtained by processing (e.g., re-encoding, re-modulating, and so on) the decoded data from RX data processor 162. The modified Yamamoto metric is indicative of the confidence in the decoded data and is obtained based on the difference

between the selected (best) path through the trellis for the convolutional decoding and the next closest path through the trellis. The minimum or average LLR may also be used as an indication of the confidence of the decoded data. These decoder metrics, which are indicative of the quality of the data transmission received on parallel channel x, are known in the art.

[1037] Outer loop 220 can provide different types of information used to control the operation of inner loop 210. For example, outer loop 220 can provide an SNR offset for each parallel channel. The SNR offset is used in the computation of the operating SNR for the parallel channel, as described below. The operating SNR is then provided to a look-up table (LUT) 176 and used to select the transmission mode for the parallel channel. The SNR offset thus influences the selection of the transmission mode. Outer loop 220 can also provide a transmission mode adjustment for each parallel channel. This adjustment may direct inner loop 210 to select a transmission mode with a lower data rate for the parallel channel. The transmission mode adjustment directly impacts the selection of the transmission mode. The SNR offset and transmission mode adjustment are two mechanisms for controlling the operation of inner loop 210. Outer loop 220 may also be designed to provide other types of adjustments for inner loop 210. For simplicity, only the SNR offset and transmission mode adjustment are described below. Outer loop 220 may adjust the SNR offset and/or transmission mode in various manners, some of which are described below.

[1038] In a first embodiment, the SNR offset and/or transmission mode for each parallel channel are adjusted based on packet errors detected for the data stream received on that parallel channel. The data stream may be transmitted in packets, blocks, frames, or some other data units. (For simplicity, packet is used herein for the data unit.) Each packet may be coded with an error detection code (e.g., a cyclic redundancy check (CRC) code) that allows the receiver to determine whether the packet was decoded correctly or in error. Each parallel channel may be associated with a particular target packet error rate (PER) (e.g., 1% PER). Quality estimator 172 receives the status of each received packet and the target PER for parallel channel x and adjusts the SNR offset for parallel channel x accordingly. For example, the SNR offset for parallel channel x may be initialized to zero at the start of data transmission on parallel channel x. The SNR offset may thereafter be reduced by ΔDN for each good packet and increased by ΔUP for each bad packet, where ΔDN and ΔUP may be selected based on

the target PER and the desired response time for the outer loop. The SNR offset is typically a positive value or zero but may also be allowed to be a negative value (e.g., to account for a high initial estimate of the received SNR). Alternatively or additionally, quality estimator 172 may provide a directive to adjust the transmission mode for parallel channel x to the next lower data rate, for example, if a burst of packet errors is detected on parallel channel x. The SNR offset and/or transmission mode adjustment from quality estimator 172 are used by selector 174 to select the transmission mode for parallel channel x.

[1039] In a second embodiment, the SNR offset and/or transmission mode for each parallel channel are adjusted based on the decoder metrics for that parallel channel. The decoder metrics for each parallel channel can be used to estimate the quality of the data transmission received on that parallel channel. If a particular decoder metric for a given parallel channel is worse than a threshold selected for that metric, then the SNR offset and/or transmission mode for that parallel channel may be adjusted accordingly.

[1040] In a third embodiment, the SNR offset and/or transmission mode for each parallel channel are adjusted based on the received SNR and the required SNR for that parallel channel. The received SNR for each parallel channel may be determined based on the received pilot symbols for that parallel channel. The system may support a set of transmission modes (e.g., as shown in Table 2), and each supported transmission mode requires a different minimum SNR to achieve the target PER. Quality estimator 172 can determine an SNR margin for parallel channel x, which is the difference between the received SNR and the required SNR for parallel channel x. If the SNR margin for parallel channel x is a negative value, then the transmission mode for parallel channel x may be adjusted to the next lower data rate.

[1041] The third embodiment may also be used for a design whereby a packet is demultiplexed and transmitted across multiple parallel channels. If the packet is received in error, then it may not be possible to determine (just from the received packet) which one or ones of the parallel channels cause the packet to be received in error. If no other information is available, then it may be necessary to adjust the N_C SNR offsets and/or the N_C transmission modes for all N_C parallel channels, for example, so that the next lower data rate is used for each parallel channel. This may result in an excessive amount of reduction on the overall data rate. However, using the third embodiment, the parallel channel with the smallest SNR margin can be assumed to have

caused the packet error, and the transmission mode for this parallel channel can be adjusted to the next lower data rate.

[1042] The outer loop may also adjust the operation of the inner loop in other manners, and this is within the scope of the invention. In general, the outer loop operates at a rate that may be faster or slower than the rate of the inner loop. For example, the adjustment of the SNR offset by the outer loop may be dependent on many received packets. The outer loop can also adjust the data rate in between regularly scheduled inner loop calculations. Thus, depending on its specific design and manner of operation, the outer loop typically has more influence on the operation of the inner loop for longer data transmissions. For bursty transmissions, the outer loop may not have much or any influence on the operation of the inner loop.

[1043] **FIG. 3** shows a flow diagram of a process 300 to transmit N_C data streams on N_C parallel channels using N_C transmission modes selected with closed-loop rate control. Process 300 may be implemented as shown in FIGS. 1 and 2. Initially, the receiver estimates the channel gains and the noise floor N_0 for the N_C parallel channels (step 312). The receiver then selects a transmission mode for each of the N_C parallel channels based on the channel gain estimates, the noise floor estimate, and outer loop information (if any) for that parallel channel (step 314). The outer loop information may include the SNR offset and/or transmission mode adjustment for each of the N_C parallel channels. The transmission mode selection is described below. The receiver sends the N_C selected transmission modes for the N_C parallel channels, as feedback information, to the transmitter (step 316).

[1044] The transmitter codes and modulates the N_C data streams in accordance with the N_C selected transmission modes (obtained from the receiver) to provide N_C modulation symbol streams (step 322). The transmitter then processes and transmits the N_C modulation symbol streams on the N_C parallel channels to the receiver (step 324).

[1045] The receiver processes the data transmissions received on the N_C parallel channels from the transmitter and obtains N_C recovered symbol streams (step 332). The receiver further processes the N_C recovered symbol streams in accordance with the N_C selected transmission modes to obtain N_C decoded data streams (step 334). The receiver also estimates the quality of the data transmission received on each of the N_C parallel channels, e.g., based on the packet status, decoder metrics, received SNRs, and so on (step 336). The receiver then provides outer loop information for each of the N_C parallel

channels based on the estimated quality for the data transmission received on that parallel channel (step 338). In FIG. 3, steps 312 through 324 may be considered as part of the inner loop, and steps 332 through 338 may be considered as part of the outer loop.

[1046] **FIG. 4** shows a flow diagram of a process 400 that may be performed for the outer loop. The status of data packets received on each of the N_C parallel channels is obtained and used to adjust the SNR offset and/or transmission mode for that parallel channel (step 412). Decoder metrics for each of the N_C parallel channels may also be obtained and used to adjust the SNR offset and/or transmission mode for that parallel channel (step 414). The received SNR for each of the N_C parallel channels may also be obtained for each parallel channel and used to compute the SNR margin for that parallel channel. The SNR margins for the N_C parallel channels may be used to adjust the transmission modes for the parallel channels if packet errors are detected (step 416). An outer loop may implement any one or any combination of the steps shown in FIG. 4, depending on its specific design.

[1047] The closed-loop rate control techniques described herein may be used for various types of multi-channel communication systems having multiple parallel channels that may be used for data transmission. For example, these techniques may be used for TDM systems, FDM systems, OFDM-based systems, MIMO systems, MIMO systems that utilize OFDM (i.e., MIMO-OFDM systems), and so on.

[1048] A TDM system may transmit data in frames, each of which may be of a particular time duration. Each frame may include multiple (N_{TS}) slots that may be assigned different slot indices. N_C parallel channels may be formed by the N_{TS} slots in each frame, where $N_C \le N_{TS}$. Each of the N_C parallel channels may include one or multiple slots. The N_C channels are considered "parallel" even though they are not transmitted simultaneously.

[1049] An FDM system may transmit data in (N_{SB}) frequency subbands, which may be arbitrarily spaced. N_C parallel channels may be formed by the N_{SB} subbands, where $N_C \le N_{SB}$. Each of the N_C parallel channels may include one or multiple subbands.

[1050] An OFDM system uses OFDM to effectively partition the overall system bandwidth into multiple (N_F) orthogonal subbands, which may also be referred to as tones, bins, and frequency channels. Each subband is associated with a respective carrier that may be modulated with data. N_C parallel channels may be formed by the N_F

subbands, where $N_C \leq N_F$. The N_C parallel channels are formed by N_C disjoints sets of one or more subbands. The N_C sets are disjoint in that each of the N_F subbands is assigned to only one set (and thus to one parallel channel), if at all. An OFDM system may be considered as a specific type of FDM system.

[1051] A MIMO system employs multiple (N_T) transmit antennas and multiple (N_R) receive antennas for data transmission, and is denoted as an (N_T, N_R) system. A MIMO channel formed by the N_T transmit and N_R receive antennas is composed of N_S spatial channels that may be used for data transmission, where $N_S \leq \min\{N_T, N_R\}$. The number of spatial channels is determined by a channel response matrix $\underline{\mathbf{H}}$ that describes the response between the N_T transmit and N_R receive antennas. For simplicity, the following description assumes that the channel response matrix $\underline{\mathbf{H}}$ is full rank. In this case, the number of spatial channels is given as $N_S = N_T \leq N_R$. N_C parallel channels may be formed by the N_S spatial channels, where $N_C \leq N_S$. Each of the N_C parallel channels may include one or multiple spatial channels.

[1052] A MIMO-OFDM system has N_S spatial channels for each of N_F subbands. N_C parallel channels may be formed by the N_S spatial channels of each of the N_F subbands, where $N_C \leq N_F \cdot N_S$. Each of the N_C parallel channels may include one or multiple spatial channels of one or multiple subbands (i.e., any combination of spatial channels and subbands). For MIMO and MIMO-OFDM systems, N_C parallel channels may also be formed by the N_T transmit antennas, where $N_C \leq N_T$. Each of the N_C parallel channels may be associated with one or multiple transmit antennas for data transmission.

[1053] For MIMO and MIMO-OFDM systems, data may be transmitted on the N_S spatial channels in various manners. For a partial channel state information (partial-CSI) MIMO system, data is transmitted on the N_S spatial channels without any spatial processing at the transmitter and with spatial processing at the receiver. For a full-CSI MIMO system, data is transmitted on the N_S spatial channels with spatial processing at both the transmitter and the receiver. For the full-CSI MIMO system, eigenvalue decomposition or singular value decomposition may be performed on the channel response matrix $\underline{\mathbf{H}}$ to obtain N_S "eigenmodes" of the MIMO channel. Data is transmitted on the N_S eigenmodes, which are orthogonalized spatial channels.

[1054] The closed-loop rate control techniques described herein may be used for time division duplex (TDD) systems as well as frequency division duplex (FDD) systems. For a TDD system, the downlink and uplink share the same frequency band and are likely to observe similar fading and multipath effects. Thus, the channel response for each link may be estimated based on a pilot received on either that link or the other link. For an FDD system, the downlink and uplink use different frequency bands and are likely to observe different fading and multipath effects. The channel response for each link may be estimated based on a pilot received on that link.

[1055] The closed-loop rate control techniques may be used for both partial-CSI and full-CSI MIMO systems. These techniques may also be used for the downlink as well as the uplink.

[1056] The closed-loop rate control techniques are now described in detail for an exemplary multi-channel communication system, which is a full-CSI TDD MIMO-OFDM system. For simplicity, in the following description, the term "eigenmode" and "wideband eigenmode" are used to denote the case where an attempt is made to orthogonalize the spatial channels, even though it may not be fully successful due to, for example, an imperfect channel estimate.

I. TDD MIMO-OFDM System

[1057] **FIG. 5** shows an exemplary TDD MIMO-OFDM system 500 with a number of access points (APs) 510 that support communication for a number of user terminals (UTs) 520. For simplicity, only two access points 510a and 510b are shown in FIG. 5. An access point may also be referred to as a base station, a base transceiver system, a Node B, or some other terminology. A user terminal may be fixed or mobile, and may also be referred to as an access terminal, a mobile station, a user equipment (UE), a wireless device, or some other terminology. Each user terminal may communicate with one or possibly multiple access points on the downlink and/or the uplink at any given moment. A system controller 530 couples to access points 510 and provides coordination and control for these access points.

[1058] **FIG. 6** shows an exemplary frame structure 600 that may be used in TDD MIMO-OFDM system 500. Data transmission occurs in units of TDD frames, each of which spans a particular time duration (e.g., 2 msec). Each TDD frame is partitioned into a downlink phase and an uplink phase, and each phase is further partitioned into

multiple segments for multiple transport channels. In the embodiment shown in FIG. 6, the downlink transport channels include a broadcast channel (BCH), a forward control channel (FCCH), and a forward channel (FCH), and the uplink transport channels include a reverse channel (RCH) and a random access channel (RACH).

[1059] In the downlink phase, a BCH segment 610 is used to transmit one BCH protocol data unit (PDU) 612, which includes a beacon pilot 614, a MIMO pilot 616, and a BCH message 618. The beacon pilot is a pilot transmitted from all antennas and is used for timing and frequency acquisition. The MIMO pilot is a pilot transmitted from all antennas but with a different orthogonal code for each antenna in order to allow the user terminals to individually identify the antennas. The MIMO pilot is used for channel estimation. The BCH message carries system parameters for the user terminals. An FCCH segment 620 is used to transmit one FCCH PDU, which carries assignments for downlink and uplink resources (e.g., the selected transmission modes for the downlink and uplink) and other signaling for the user terminals. An FCH segment 630 is used to transmit one or more FCH PDUs 632 on the downlink. Different types of FCH PDU may be defined. For example, an FCH PDU 632a includes a steered reference 634a and a data packet 636a, and an FCH PDU 632b includes only a data packet 636b. The steered reference is a pilot that is transmitted on a specific wideband eigenmode (as described below) and is used for channel estimation.

[1060] In the uplink phase, an RCH segment 640 is used to transmit one or more RCH PDUs 642 on the uplink. Different types of RCH PDU may also be defined. For example, an RCH PDU 642a includes only a data packet 646a, and an RCH PDU 642b includes a steered reference 644b and a data packet 646b. An RACH segment 650 is used by the user terminals to gain access to the system and to send short messages on the uplink. An RACH PDU 652 may be sent in RACH segment 650 and includes a pilot (e.g., steered reference) 654 and a message 656.

[1061] FIG. 6 shows an exemplary frame structure for a TDD system. Other frame structures may also be used, and this is within the scope of the invention.

1. Spatial Processing

[1062] For a MIMO-OFDM system, the channel response between an access point and a user terminal may be characterized by a set of channel response matrices, $\underline{\mathbf{H}}(k)$ for $k \in K$, where K represents the set of all subbands of interest (e.g., $K = \{1, ..., N_F\}$).

For a TDD MIMO-OFDM system with a shared frequency band, the downlink and uplink channel responses may be assumed to be reciprocal of one another. That is, if $\underline{\mathbf{H}}(k)$ represents a channel response matrix from antenna array A to antenna array B for subband k, then a reciprocal channel implies that the coupling from array B to array A is given by $\underline{\mathbf{H}}^T(k)$, where $\underline{\mathbf{A}}^T$ denotes the transpose of $\underline{\mathbf{A}}$.

[1063] However, the frequency responses of the transmit and receive chains at the access point are typically different from the frequency responses of the transmit and receive chains at the user terminal. Calibration may be performed to obtain correction matrices used to account for differences in the frequency responses. With these correction matrices, the "calibrated" downlink channel response, $\underline{\mathbf{H}}_{\text{cdn}}(k)$, observed by the user terminal is the transpose of the "calibrated" uplink channel response, $\underline{\mathbf{H}}_{\text{cup}}(k)$, observed by the access point, i.e., $\underline{\mathbf{H}}_{\text{cdn}}(k) = \underline{\mathbf{H}}_{\text{cup}}^T(k)$, for $k \in K$. For simplicity, the following description assumes that the downlink and uplink channel responses are calibrated and reciprocal of one another.

[1064] On the downlink, a MIMO pilot may be transmitted by the access point (e.g., in BCH segment 610) and used by the user terminal to obtain an estimate of the calibrated downlink channel response, $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$, for $k \in K$. The user terminal may estimate the calibrated uplink channel response as $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k) = \hat{\underline{\mathbf{H}}}_{\text{cdn}}^T(k)$. The user terminal may perform singular value decomposition of $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$, for each subband k, as follows:

$$\hat{\underline{\mathbf{H}}}_{\text{cup}}(k) = \hat{\underline{\mathbf{U}}}_{\text{ap}}(k)\hat{\underline{\Sigma}}(k)\hat{\underline{\mathbf{V}}}_{\text{ut}}^H(k) \quad \text{, for } k \in K, \qquad \text{Eq (1)}$$

where $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ is an $(N_{ap} \times N_{ap})$ unitary matrix of left eigenvectors of $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$;

$\hat{\underline{\Sigma}}(k)$ is an $(N_{ap} \times N_{ut})$ diagonal matrix of singular values of $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$;

$\hat{\underline{\mathbf{V}}}_{\text{ut}}(k)$ is an $(N_{ut} \times N_{ut})$ unitary matrix of right eigenvectors of $\hat{\underline{\mathbf{H}}}_{\text{cup}}(k)$;

$\underline{\mathbf{A}}^H$ is the conjugate transpose of $\underline{\mathbf{A}}$;

N_{ap} is the number of antennas at the access point; and

N_{ut} is the number of antennas at the user terminal.

[1065] Similarly, the singular value decomposition of $\hat{\underline{\mathbf{H}}}_{\text{cdn}}(k)$ may be expressed as:

$$\underline{\hat{\mathbf{H}}}_{\text{cdn}}(k) = \underline{\hat{\mathbf{V}}}_{\text{ut}}^{*}(k)\underline{\hat{\Sigma}}(k)\underline{\hat{\mathbf{U}}}_{\text{ap}}^{T}(k) \quad \text{, for } k \in K \text{,} \qquad \text{Eq (2)}$$

where $\underline{\hat{\mathbf{V}}}_{\text{ut}}^{*}(k)$ and $\underline{\hat{\mathbf{U}}}_{\text{ap}}^{*}(k)$ are unitary matrices of left and right eigenvectors, respectively, of $\underline{\hat{\mathbf{H}}}_{\text{cdn}}(k)$ and "*" denotes the complex conjugate. Singular value decomposition is described by Gilbert Strang in a book entitled "Linear Algebra and Its Applications," Second Edition, Academic Press, 1980.

[1066] As shown in equations (1) and (2), the matrices of left and right eigenvectors for one link are the complex conjugate of the matrices of right and left eigenvectors, respectively, for the other link. The matrices $\underline{\hat{\mathbf{U}}}_{\text{ap}}(k)$ and $\underline{\hat{\mathbf{V}}}_{\text{ut}}(k)$ may be used by the access point and the user terminal, respectively, for spatial processing and are denoted as such by their subscripts. The matrix $\underline{\hat{\Sigma}}(k)$ includes singular value estimates that represent the gains for the spatial channels (or eigenmodes) of the channel response matrix $\underline{\mathbf{H}}(k)$ for each subband k.

[1067] Singular value decomposition may be performed independently for the channel response matrix $\underline{\hat{\mathbf{H}}}_{\text{cup}}(k)$ for each subband k to determine the N_S eigenmodes of that subband. The singular value estimates for each diagonal matrix $\underline{\hat{\Sigma}}(k)$ may be ordered such that $\{\hat{\sigma}_1(k) \geq \hat{\sigma}_2(k) \geq \ldots \geq \hat{\sigma}_{N_S}(k)\}$, where $\hat{\sigma}_1(k)$ is the largest singular value estimate and $\hat{\sigma}_{N_S}(k)$ is the smallest singular value estimate for subband k. When the singular value estimates for each diagonal matrix $\underline{\hat{\Sigma}}(k)$ are ordered, the eigenvectors (or columns) of the associated matrices $\underline{\hat{\mathbf{U}}}(k)$ and $\underline{\hat{\mathbf{V}}}(k)$ are also ordered correspondingly. A "wideband eigenmode" may be defined as the set of same-order eigenmodes of all subbands after the ordering. Thus, the m-th wideband eigenmode includes the m-th eigenmodes of all subbands. The "principal" wideband eigenmode is the one associated with the largest singular value estimate in the matrix $\underline{\hat{\Sigma}}(k)$ for each of the subbands. N_S parallel channels may be formed by the N_S wideband eigenmodes.

[1068] The user terminal may transmit a steered reference on the uplink (e.g., in RCH segment 640 or RACH segment 650 in FIG. 6). The uplink steered reference for wideband eigenmode m may be expressed as:

$$\underline{\mathbf{x}}_{\text{up,sr},m}(k) = \underline{\hat{\mathbf{v}}}_{\text{ut},m}(k)p(k) \quad \text{, for } k \in K \text{,} \qquad \text{Eq (3)}$$

where $\underline{\mathbf{x}}_{\mathrm{up,sr},m}(k)$ is a vector of N_{ut} symbols sent from N_{ut} user terminal antennas for subband k of wideband eigenmode m for the steered reference;

$\hat{\underline{\mathbf{v}}}_{\mathrm{ut},m}(k)$ is the m-th column of the matrix $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$ for subband k, where $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k) = [\hat{\underline{\mathbf{v}}}_{\mathrm{ut},1}(k)\ \hat{\underline{\mathbf{v}}}_{\mathrm{ut},2}(k)\ \dots\ \hat{\underline{\mathbf{v}}}_{\mathrm{ut},N_{ut}}(k)]$; and

$p(k)$ is the pilot symbol sent on subband k.

The steered reference for all N_S wideband eigenmodes may be transmitted in N_S OFDM symbol periods, or fewer than N_S OFDM symbol periods using subband multiplexing. The steered reference for each wideband eigenmode may also be transmitted over multiple OFDM symbol periods.

[1069] The received uplink steered reference at the access point may be expressed as:

$$\begin{aligned}\underline{\mathbf{r}}_{\mathrm{up,sr},m}(k) &= \underline{\mathbf{H}}_{\mathrm{cup}}(k)\hat{\underline{\mathbf{v}}}_{\mathrm{ut},m}(k)p(k) + \underline{\mathbf{n}}_{\mathrm{up}}(k) \\ &\approx \hat{\underline{\mathbf{u}}}_{\mathrm{ap},m}(k)\hat{\sigma}_m(k)p(k) + \underline{\mathbf{n}}_{\mathrm{up}}(k)\end{aligned}\ , \text{ for } k \in K\,, \qquad \text{Eq (4)}$$

where $\underline{\mathbf{r}}_{\mathrm{up,sr},m}(k)$ is a vector of N_{ap} symbols received on N_{ap} access point antennas for subband k of wideband eigenmode m for the steered reference;

$\hat{\underline{\mathbf{u}}}_{\mathrm{ap},m}(k)$ is the m-th column of the matrix $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k)$ for subband k, where $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}(k) = [\hat{\underline{\mathbf{u}}}_{\mathrm{ap},1}(k)\ \hat{\underline{\mathbf{u}}}_{\mathrm{ap},2}(k)\ \dots\ \hat{\underline{\mathbf{u}}}_{\mathrm{ap},N_{ap}}(k)]$;

$\hat{\sigma}_m(k)$ is the singular value estimate for subband k of wideband eigenmode m, i.e., the m-th diagonal element of the matrix $\hat{\underline{\Sigma}}(k)$; and

$\underline{\mathbf{n}}_{\mathrm{up}}(k)$ is additive white Gaussian noise (AWGN) for subband k on the uplink.

[1070] As shown in equation (4), at the access point, the received steered reference (in the absence of noise) is approximately $\hat{\underline{\mathbf{u}}}_{\mathrm{ap},m}(k)\hat{\sigma}_m(k)p(k)$. The access point can thus obtain estimates of both $\hat{\underline{\mathbf{u}}}_{\mathrm{ap},m}(k)$ and $\hat{\sigma}_m(k)$ for each subband k based on the received steered reference for that subband. The estimate of $\hat{\sigma}_m(k)$ for subband k of wideband eigenmode m, $\hat{\hat{\sigma}}_m(k)$, may be expressed as:

$$\hat{\hat{\sigma}}_m(k) = \| \underline{\mathbf{r}}_{\text{up,sr},m}(k) \|^2 = \sum_{i=1}^{N_{ap}} | r_{\text{up,sr},m,i}(k) |^2 \quad , \text{ for } k \in K \text{ and } m \in M , \qquad \text{Eq (5)}$$

where $\| \underline{\mathbf{a}} \|$ denotes the 2-norm of $\underline{\mathbf{a}}$;

$r_{\text{up,sr},m,i}(k)$ is the i-th element of the vector $\underline{\mathbf{r}}_{\text{up,sr},m}(k)$; and

M represents the set of all wideband eigenmodes of interest, e.g., $M = \{1, ..., N_S\}$.

[1071] The estimate of $\hat{\underline{\mathbf{u}}}_{\text{ap},m}(k)$ for subband k of wideband eigenmode m, $\hat{\hat{\underline{\mathbf{u}}}}_{\text{ap},m}(k)$, may be expressed as:

$$\hat{\hat{\underline{\mathbf{u}}}}_{\text{ap},m}(k) = \underline{\mathbf{r}}_{\text{up,sr},m}(k) / \hat{\hat{\sigma}}_m(k) \quad , \text{ for } k \in K \text{ and } m \in M . \qquad \text{Eq (6)}$$

The double hat for $\hat{\hat{\underline{\mathbf{u}}}}_{\text{ap},m}(k)$ and $\hat{\hat{\sigma}}_m(k)$ indicates that these are estimates of estimates, i.e., estimates obtained by the access point for the estimates $\hat{\underline{\mathbf{u}}}_{\text{ap},m}(k)$ and $\hat{\sigma}_m(k)$ obtained by the user terminal. If the steered reference for each wideband eigenmode is transmitted over multiple OFDM symbol periods, then the access point can average the received steered reference for each wideband eigenmode to obtain more accurate estimates of $\hat{\underline{\mathbf{u}}}_{\text{ap},m}(k)$ and $\hat{\sigma}_m(k)$.

[1072] Table 1 summarizes the spatial processing at the access point and the user terminal for data transmission and reception on multiple wideband eigenmodes.

Table 1

	Downlink	**Uplink**
Access Point	Transmit : $\underline{\mathbf{x}}_{\text{dn}}(k) = \hat{\hat{\underline{\mathbf{U}}}}{}^{*}_{\text{ap}}(k)\underline{\mathbf{s}}_{\text{dn}}(k)$	Receive : $\hat{\underline{\mathbf{s}}}_{\text{up}}(k) = \hat{\hat{\underline{\Sigma}}}{}^{-1}(k)\hat{\hat{\underline{\mathbf{U}}}}{}^{H}_{\text{ap}}(k)\underline{\mathbf{r}}_{\text{up}}(k)$
User Terminal	Receive : $\hat{\underline{\mathbf{s}}}_{\text{dn}}(k) = \hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{V}}}{}^{T}_{\text{ut}}(k)\underline{\mathbf{r}}_{\text{dn}}(k)$	Transmit : $\underline{\mathbf{x}}_{\text{up}}(k) = \hat{\underline{\mathbf{V}}}_{\text{ut}}(k)\underline{\mathbf{s}}_{\text{up}}(k)$

In Table 1, $\underline{\mathbf{s}}(k)$ is a "data" vector of modulation symbols (obtained from the symbol mapping at the transmitter), $\underline{\mathbf{x}}(k)$ is a "transmit" vector of transmit symbols (obtained after spatial processing at the transmitter), $\underline{\mathbf{r}}(k)$ is a "received" vector of received symbols (obtained after OFDM processing at the receiver), and $\hat{\underline{\mathbf{s}}}(k)$ is an estimate of

the vector $\underline{\mathbf{s}}(k)$ (obtained after spatial processing at the receiver), where all of the vectors are for subband k. The subscripts "dn" and "up" for these vectors denote downlink and uplink, respectively. In Table 1, $\underline{\Sigma}^{-1}(k)$ is a diagonal matrix defined as $\underline{\Sigma}^{-1}(k) = \text{diag}\ (1/\sigma_1(k)\ 1/\sigma_2(k)\ \ldots\ 1/\sigma_{N_S}(k))$.

[1073] The steered reference may be transmitted for one wideband eigenmode at a time by the user terminal or may be transmitted for multiple wideband eigenmodes simultaneously using an orthogonal basis (e.g., Walsh codes). The steered reference for each wideband eigenmode may be used by the access point to obtain $\hat{\underline{\mathbf{u}}}_{\text{ap},m}(k)$, for $k \in K$, for that wideband eigenmode. If the N_S vectors $\hat{\underline{\mathbf{u}}}_{\text{ap},m}(k)$ of the matrix $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ are obtained individually (and over different OFDM symbol periods) for the N_S eigenmodes of each subband, then, due to noise and other sources of degradation in the wireless link, the N_S vectors $\hat{\underline{\mathbf{u}}}_{\text{ap},m}(k)$ of the matrix $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ for each subband k are not likely to be orthogonal to one another. In this case, the N_S vectors of the matrix $\hat{\underline{\mathbf{U}}}_{\text{ap}}(k)$ for each subband k may be orthogonalized using QR factorization, polar decomposition, or some other techniques.

[1074] At the access point, a received SNR estimate for subband k of wideband eigenmode m, $\gamma_{\text{ap},m}(k)$, may be expressed as:

$$\gamma_{\text{ap},m}(k) = \frac{P_{\text{up},m}(k) \cdot \hat{\sigma}_m^2(k)}{N_{0,\text{ap}}}\ ,\ \text{for}\ k \in K\ \text{and}\ m \in M\ , \qquad \text{Eq (7)}$$

where $P_{\text{up},m}(k)$ is the transmit power used by the user terminal for subband k of wideband eigenmode m on the uplink; and

$N_{0,\text{ap}}$ is the noise floor at the access point.

[1075] At the user terminal, a received SNR estimate for subband k of wideband eigenmode m, $\gamma_{\text{ut},m}(k)$, may be expressed as:

$$\gamma_{\text{ut},m}(k) = \frac{P_{\text{dn},m}(k) \cdot \hat{\sigma}_m^2(k)}{N_{0,\text{ut}}}\ ,\ \text{for}\ k \in K\ \text{and}\ m \in M\ , \qquad \text{Eq (8)}$$

where $P_{dn,m}(k)$ is the transmit power used by the access point for subband k of wideband eigenmode m on the downlink; and

$N_{0,ut}$ is the noise floor at the user terminal.

As shown in equations (7) and (8), the received SNR for each subband of each wideband eigenmode, $\gamma_m(k)$, is dependent on the channel gain (which is $\hat{\sigma}_m(k)$ or $\hat{\hat{\sigma}}_m(k)$), the receiver noise floor N_0, and the transmit power $P_m(k)$. The received SNR may be different for different subbands and eigenmodes.

[1076] **FIG. 7** shows a flow diagram of a process 700 for transmitting multiple data streams on multiple wideband eigenmodes on the downlink and uplink in the exemplary TDD MIMO-OFDM system. Process 700 assumes that calibration has already been performed and that the downlink and uplink channel responses are transpose of one another, i.e., $\hat{\underline{\mathbf{H}}}_{cup}(k) \approx \hat{\underline{\mathbf{H}}}_{cdn}^T(k)$. For process 700, channel estimation is performed in block 710, transmission mode selection is performed in block 730, and data transmission/reception is performed in block 760.

[1077] For channel estimation, the access point transmits a MIMO pilot on the downlink (e.g., on the BCH) (step 712). The user terminal receives and processes the MIMO pilot to obtain an estimate of the calibrated downlink channel response, $\hat{\underline{\mathbf{H}}}_{cdn}(k)$, for $k \in K$ (step 714). The user terminal then estimates the calibrated uplink channel response as $\hat{\underline{\mathbf{H}}}_{cup}(k) = \hat{\underline{\mathbf{H}}}_{cdn}^T(k)$ and performs singular value decomposition (SVD) of $\hat{\underline{\mathbf{H}}}_{cup}(k)$ to obtain the matrices $\hat{\underline{\Sigma}}(k)$ and $\hat{\underline{\mathbf{V}}}_{ut}(k)$, for $k \in K$, as shown in equation (1) (step 716). The user terminal then transmits an uplink steered reference (e.g., on the RACH or the RCH) using the matrices $\hat{\underline{\mathbf{V}}}_{ut}(k)$, for $k \in K$, as shown in equation (3) (step 718). The access point receives and processes the uplink steered reference to obtain the matrices $\hat{\hat{\underline{\Sigma}}}(k)$ and $\hat{\hat{\underline{\mathbf{U}}}}_{ap}(k)$, for $k \in K$, as described above (step 720).

[1078] For downlink data transmission, the user terminal selects a transmission mode (with the highest supported data rate) for each wideband eigenmode on the downlink based on the diagonal matrix $\hat{\underline{\Sigma}}(k)$, the noise floor $N_{0,ut}$ at the user terminal, and downlink outer loop information (e.g., SNR offsets and/or transmission mode adjustments for the downlink) (step 740). The transmission mode selection is described

below. The user terminal then sends feedback information, which includes the N_S transmission modes selected by the user terminal for the downlink and may further include the noise floor $N_{0,\mathrm{ut}}$ at the user terminal (step 742). (The steered reference transmitted in step 718 may also be viewed as feedback information sent by the user terminal.)

[1079] For uplink data transmission, the access point selects N_S transmission modes for the N_S wideband eigenmodes on the uplink based on the diagonal matrix $\hat{\underline{\Sigma}}(k)$, the noise floor $N_{0,\mathrm{ap}}$ at the access point, and uplink outer loop information (e.g., SNR offsets and/or transmission mode adjustments for the uplink) (step 750). The access point further selects the N_S transmission modes for the N_S wideband eigenmodes on the downlink based on the feedback information received from the user terminal (step 752). The access point then sends the selected transmission modes for both the downlink and uplink (e.g., on the FCCH) (step 754). The user terminal receives the selected transmission modes for both links (step 756).

[1080] For downlink data transmission, the access point (1) codes and modulates the data for each downlink wideband eigenmode in accordance with the transmission mode selected for that wideband eigenmode, (2) spatially processes the data vector $\underline{\mathbf{s}}_{\mathrm{dn}}(k)$ with the matrix $\hat{\underline{\mathbf{U}}}_{\mathrm{ap}}^{*}(k)$, as shown in Table 1, to obtain the transmit vector $\underline{\mathbf{x}}_{\mathrm{dn}}(k)$, for $k \in K$, and (3) transmits the vector $\underline{\mathbf{x}}_{\mathrm{dn}}(k)$ on the downlink (step 762). The user terminal (1) receives the downlink transmission, (2) performs matched filtering on the received vector $\underline{\mathbf{r}}_{\mathrm{dn}}(k)$ with $\hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}^{T}(k)$, as also shown in Table 1, to obtain the vector $\hat{\underline{\mathbf{s}}}_{\mathrm{dn}}(k)$, for $k \in K$, and (3) demodulates and decodes the recovered symbols in accordance with the transmission mode selected for each downlink wideband eigenmode (step 764).

[1081] For uplink data transmission, the user terminal (1) codes and modulates the data for each uplink wideband eigenmode in accordance with the transmission mode selected for that wideband eigenmode, (2) spatially processes the data vector $\underline{\mathbf{s}}_{\mathrm{up}}(k)$ with the matrix $\hat{\underline{\mathbf{V}}}_{\mathrm{ut}}(k)$ to obtain the transmit vector $\underline{\mathbf{x}}_{\mathrm{up}}(k)$, for $k \in K$, and (3) transmits the vector $\underline{\mathbf{x}}_{\mathrm{up}}(k)$ on the uplink (step 772). The access point (1) receives the uplink transmission, (2) performs matched filtering on the received vector $\underline{\mathbf{r}}_{\mathrm{up}}(k)$ with

$\hat{\underline{\Sigma}}^{-1}(k)\hat{\underline{\mathbf{U}}}_{ap}^{H}(k)$ to obtain the vector $\hat{\underline{\mathbf{s}}}_{up}(k)$, and (3) demodulates and decodes the recovered symbols in accordance with the transmission mode selected for each uplink wideband eigenmode (step 774). For simplicity, the closed-loop operation and the transmission mode adjustment by the outer loop are not shown in FIG. 7.

[1082] FIG. 7 shows a specific embodiment of a process that may be used for downlink and uplink data transmission in the exemplary TDD MIMO-OFDM system. Other processes may also be implemented whereby the channel estimation, transmission mode selection, and/or data transmission/reception may be performed in some other manners.

2. Transmission Mode Selection

[1083] **FIG. 8** shows a flow diagram of a process 800 for selecting N_S transmission modes for the N_S wideband eigenmodes. Process 800 may be used for steps 740 and 750 in FIG. 7. Initially, the total transmit power, P_{total}, available at the transmitter for data transmission is distributed to the N_S wideband eigenmodes based on a power distribution scheme (step 812). The transmit power P_m allocated to each wideband eigenmode is then distributed to the N_F subbands of that wideband eigenmode based on the same or a different power distribution scheme (step 814). The power distribution across the N_S wideband eigenmodes and the power distribution across the N_F subbands of each wideband eigenmode may be performed as described below.

[1084] An operating SNR for each wideband eigenmode, $\gamma_{op,m}$, is computed based on (1) the allocated transmit powers $P_m(k)$ and the channel gains $\sigma_m(k)$ for the subbands of that wideband eigenmode, (2) the noise floor N_0 at the receiver, and (3) the SNR offset for that wideband eigenmode (step 816). The computation of the operating SNR is described below. A suitable transmission mode q_m is then selected for each wideband eigenmode based on the operating SNR for that wideband eigenmode and a look-up table (step 818). Excess power for each wideband eigenmode is determined, and the total excess power for all wideband eigenmodes is redistributed to one or more wideband eigenmodes to improve performance (step 820). The transmission mode for each wideband eigenmode may be adjusted (e.g., to the next lower data rate) if directed by outer loop information (step 822). Each of the steps in FIG. 8 is described in detail below.

A. Power Distribution Across Wideband Eigenmodes

[1085] For step 812 in FIG. 8, the total transmit power, P_{total}, may be distributed to the N_S wideband eigenmodes using various schemes. Some of these power distribution schemes are described below.

[1086] In a uniform power distribution scheme, the total transmit power, P_{total}, is distributed uniformly across the N_S wideband eigenmodes such that they are all allocated equal power. The transmit power P_m allocated to each wideband eigenmode m may be expressed as:

$$P_m = \frac{P_{\text{total}}}{N_S} \quad , \text{ for } m \in M . \qquad \text{Eq (9)}$$

[1087] In a water-filling power distribution scheme, the total transmit power, P_{total}, is distributed to the N_S wideband eigenmodes based on a "water-filling" or "water-pouring" procedure. The water-filling procedure distributes the total transmit power, P_{total}, across the N_S wideband eigenmodes such that the overall spectral efficiency is maximized. Water-filling is described by Robert G. Gallager in "Information Theory and Reliable Communication," John Wiley and Sons, 1968. The water-filling for the N_S wideband eigenmodes may be performed in various manners, some of which are described below.

[1088] In a first embodiment, the total transmit power, P_{total}, is initially distributed to the $N_S N_F$ subbands/eigenmodes using water-filling and based on their received SNRs, $\gamma_m(k)$, for $k \in K$ and $m \in M$. The received SNR, $\gamma_m(k)$, may be computed as shown in equation (7) or (8) with the assumption of P_{total} being uniformly distributed across the $N_S N_F$ subbands/eigenmodes. The result of this power distribution is an initial transmit power, $P'_m(k)$, for each subband/eigenmode. The transmit power P_m allocated to each wideband eigenmode is then obtained by summing the initial transmit powers, $P'_m(k)$, allocated to the N_F subbands of that wideband eigenmode, as follows:

$$P_m = \sum_{k=1}^{N_F} P'_m(k) \quad , \text{ for } m \in M . \qquad \text{Eq (10)}$$

[1089] In a second embodiment, the total transmit power, P_{total}, is distributed to the N_S wideband eigenmodes based on the average SNRs computed for these wideband eigenmodes. Initially, the average SNR, $\gamma_{\text{avg},m}$, is computed for each wideband eigenmode m based on the received SNRs for the N_F subbands of that wideband eigenmode, as follows:

$$\gamma_{\text{avg},m} = \frac{1}{N_F} \sum_{k=1}^{N_F} \gamma_m(k) \ , \qquad \text{Eq (11)}$$

where $\gamma_m(k)$ is computed as described above for the first embodiment. Water-filling is then performed to distribute the total transmit power, P_{total}, across the N_S wideband eigenmodes based on their average SNRs, $\gamma_{\text{avg},m}$, for $m \in M$.

[1090] In a third embodiment, the total transmit power, P_{total}, is distributed to the N_S wideband eigenmodes based on the average SNRs for these wideband eigenmodes after channel inversion is applied for each wideband eigenmode. For this embodiment, the total transmit power, P_{total}, is first distributed uniformly to the N_S wideband eigenmodes. Channel inversion is then performed (as described below) independently for each wideband eigenmode to determine an initial power allocation, $P_m''(k)$, for each subband of that wideband eigenmode. After the channel inversion, the received SNR is the same across all subbands of each wideband eigenmode. The average SNR for each wideband eigenmode is then equal to the received SNR for any one of the subbands of that wideband eigenmode. The received SNR, $\gamma_m''(k)$, for one subband of each wideband eigenmode can be determined based on the initial power allocation, $P_m''(k)$, as shown in equation (7) or (8). The total transmit power, P_{total}, is then distributed to the N_S wideband eigenmodes using water-filling and based on their average SNRs, $\gamma''_{\text{avg},m}$, for $m \in M$.

[1091] Other schemes may also be used to distribute the total transmit power to the N_S wideband eigenmodes, and this is within the scope of the invention.

B. Power Allocation Across Subbands in Each Wideband Eigenmode

[1092] For step 814 in FIG. 8, the transmit power allocated to each wideband eigenmode, P_m, may be distributed to the N_F subbands of that wideband eigenmode using various schemes. Some of these power distribution schemes are described below.

[1093] In a uniform power distribution scheme, the transmit power for each wideband eigenmode, P_m, is distributed uniformly across the N_F subbands such that they are all allocated equal power. The transmit power $P_m(k)$ allocated to each subband may be expressed as:

$$P_m(k) = \frac{P_m}{N_F} \ , \text{ for } k \in K \text{ and } m \in M . \qquad \text{Eq (12)}$$

For the uniform power distribution scheme, the received SNRs for the N_F subbands of each wideband eigenmode are likely to vary across the subbands.

[1094] In a channel inversion scheme, the transmit power for each wideband eigenmode, P_m, is distributed non-uniformly across the N_F subbands such that they achieve similar received SNRs at the receiver. In the following description, $\sigma_m(k)$ denotes the estimated channel gain, which is equal to $\hat{\sigma}_m(k)$ for the downlink and $\hat{\hat{\sigma}}_m(k)$ for the uplink. For the channel inversion scheme, a normalization b_m is initially computed for each wideband eigenmode, as follows:

$$b_m = \frac{1}{\sum_{k=1}^{N_F} [1/\sigma_m^2(k)]} \ , \text{ for } m \in M . \qquad \text{Eq (13)}$$

The transmit power $P_m(k)$ allocated to each subband of each wideband eigenmode may then be computed as:

$$P_m(k) = \frac{b_m \cdot P_m}{\sigma_m^2(k)} \ , \text{ for } k \in K \text{ and } m \in M . \qquad \text{Eq (14)}$$

A transmit weight, $W_m(k)$, may be computed for each subband of each wideband eigenmode, as follows:

$$W_m(k) = \sqrt{P_m(k)} \ , \text{ for } k \in K \text{ and } m \in M . \qquad \text{Eq (15)}$$

The transmit weights are used to scale modulation symbols at the transmitter. For the channel inversion scheme, all N_F subbands are used for each wideband eigenmode and the received SNRs for the subbands are approximately equal.

[1095] In a selective channel inversion scheme, the transmit power for each wideband eigenmode, P_m, is distributed non-uniformly across selected ones of the N_F subbands such that the selected subbands achieve similar received SNRs at the receiver. The selected subbands are those with channel gains equal to or greater than a gain threshold. For this scheme, an average power gain, g_m, is initially computed for each wideband eigenmode, as follows:

$$g_m = \frac{1}{N_F} \sum_{k=1}^{N_F} \sigma_m^2(k) \quad , \text{ for } m \in M . \qquad \text{Eq (16)}$$

A normalization $\tilde{b}_m$ is then computed for each wideband eigenmode, as follows:

$$\tilde{b}_m = \frac{1}{\sum\limits_{\sigma_m^2(k) > \beta_m g_m} [1/\sigma_m^2(k)]} \quad , \text{ for } m \in M , \qquad \text{Eq (17)}$$

where $\beta_m g_m$ is the gain threshold and β_m is a scaling factor, which may be selected to maximize the overall throughput or based on some other criterion. The transmit power allocated to each subband of each wideband eigenmode, $P_m(k)$, may be expressed as:

$$P_m(k) = \begin{cases} \dfrac{\tilde{b}_m \cdot P_m}{\sigma_m^2(k)}, & \text{if } \sigma_m^2(k) \ge \beta_m g_m \\ 0 \ , & \text{otherwise} \end{cases} \quad , \text{ for } k \in K \text{ and } m \in M . \qquad \text{Eq (18)}$$

For the selective channel inversion scheme, N_F or fewer subbands may be selected for use for each wideband eigenmode and the received SNRs for the selected subbands are approximately equal.

[1096] Other schemes may also be used to distribute the transmit power P_m across the N_F subbands of each wideband eigenmode, and this is within the scope of the invention.

C. Transmission Mode Selection for Each Wideband Eigenmode

[1097] For step 816 in FIG. 8, an operating SNR is computed for each wideband eigenmode. The operating SNR indicates the transmission capability of the wideband eigenmode. Various methods may be used for step 816, depending on whether the received SNRs are similar or vary across the subbands of each wideband eigenmode. In the following description, SNRs are given in units of decibels (dB).

[1098] If channel inversion or selective channel inversion is performed, then the received SNRs for the subbands of each wideband eigenmode, $\gamma_m(k)$ for $k \in K$, are similar. The received SNR for subband k of wideband eigenmode m, $\gamma_m(k)$, may be computed as:

$$\gamma_m(k) = 10\log_{10}\left(\frac{P_m(k)\cdot\sigma_m^2(k)}{N_0}\right) \text{, for } k \in K \text{ and } m \in M \text{. (dB)} \qquad \text{Eq (19)}$$

The operating SNR for each wideband eigenmode, $\gamma_{\text{op},m}$, is equal to the received SNR for any one of the subbands of that wideband eigenmode minus the SNR offset for that wideband eigenmode, as follows:

$$\gamma_{\text{op},m} = \gamma_m(k) - \gamma_{\text{os},m} \text{ ,} \qquad \text{for any } k \text{ and } m \in M \text{, (dB)} \qquad \text{Eq (20)}$$

where $\gamma_m(k)$, $\gamma_{\text{os},m}$, and $\gamma_{\text{op},m}$ are all given in units of dB in equations (19) and (20).

[1099] If the transmit power P_m for each wideband eigenmode is uniformly distributed across the subbands, then the received SNRs for the subbands of each wideband eigenmode are likely to vary. In this case, the operating SNR for each wideband eigenmode, $\gamma_{\text{op},m}$, may be computed as:

$$\gamma_{\text{op},m} = \gamma_{\text{avg},m} - \gamma_{\text{bo},m} - \gamma_{\text{os},m} \text{ ,} \qquad \text{(dB)} \qquad \text{Eq (21)}$$

where $\gamma_{\text{avg},m}$ is an average of the received SNRs for the N_F subbands of wideband eigenmode m; and

$\gamma_{\text{bo},m}$ is a back-off factor that accounts for variation in the received SNRs, which may be a function of the variance of the received SNRs.

[1100] For step 818 in FIG. 8, a suitable transmission mode is selected for each wideband eigenmode based on the operating SNR for that wideband eigenmode. The system may be designed to support a set of transmission modes. The transmission mode having index 0 is for a null data rate (i.e., no data transmission). Each supported transmission mode is associated with a particular minimum SNR required to achieve the desired level of performance (e.g., 1% PER). Table 2 lists an exemplary set of 14 transmission modes supported by the system, which are identified by transmission mode indices 0 through 13. Each transmission mode is associated with a particular spectral efficiency, a particular code rate, a particular modulation scheme, and the minimum SNR required to achieve 1% PER for a non-fading, AWGN channel. The spectral efficiency refers to the data rate (i.e., the information bit rate) normalized by the system bandwidth, and is given in units of bits per second per Hertz (bps/Hz). The spectral efficiency for each transmission mode is determined by the coding scheme and the modulation scheme for that transmission mode. The code rate and modulation scheme for each transmission mode in Table 2 are specific to the exemplary system design.

Table 2

Transmission Mode Index	**Spectral Efficiency** (bps/Hz)	**Code Rate**	**Modulation Scheme**	**Required SNR** (dB)
0	0.0	-	-	-
1	0.25	1/4	BPSK	-1.8
2	0.5	1/2	BPSK	1.2
3	1.0	1/2	QPSK	4.2
4	1.5	3/4	QPSK	6.8
5	2.0	1/2	16 QAM	10.1
6	2.5	5/8	16 QAM	11.7
7	3.0	3/4	16 QAM	13.2
8	3.5	7/12	64 QAM	16.2
9	4.0	2/3	64 QAM	17.4
10	4.5	3/4	64 QAM	18.8
11	5.0	5/6	64 QAM	20.0
12	6.0	3/4	256 QAM	24.2
13	7.0	7/8	256 QAM	26.3

[1101] For each supported transmission mode with a non-zero data rate, the required SNR is obtained based on the specific system design (i.e., the particular code rate, interleaving scheme, modulation scheme, and so on, used by the system for that transmission mode) and for an AWGN channel. The required SNR may be obtained by computer simulation, empirical measurements, and so on, as is known in the art. A look-up table may be used to store the set of supported transmission modes and their required SNRs.

[1102] The operating SNR for each wideband eigenmode, $\gamma_{\mathrm{op},m}$, may be provided to the look-up table, which then provides the transmission mode q_m for that wideband eigenmode. This transmission mode q_m is the supported transmission mode with the highest data rate and a required SNR, $\gamma_{\mathrm{req},m}$, that is less than or equal to the operating SNR (i.e., $\gamma_{\mathrm{req},m} \le \gamma_{\mathrm{op},m}$). The look-up table thus selects the highest possible data rate for each wideband eigenmode based on the operating SNR for that wideband eigenmode.

D. Reallocation of Transmit Power

[1103] For step 820 in FIG. 8, the excess transmit power for each wideband eigenmode is determined and redistributed to improve performance. The following terms are used for the description below:

- Active wideband eigenmode - a wideband eigenmode with a non-zero data rate (i.e., a transmission mode having an index from 1 through 13 in Table 2);
- Saturated wideband eigenmode - a wideband eigenmode with the maximum data rate (i.e., transmission mode having index 13); and
- Unsaturated wideband eigenmode - an active wideband eigenmode with a non-zero data rate less than the maximum data rate (i.e., a transmission mode having an index from 1 through 12).

[1104] The operating SNR for a wideband eigenmode may be less than the smallest required SNR in the look-up table (i.e., $\gamma_{\mathrm{op},m} < -1.8$ dB for the transmission modes shown in Table 2). In this case, the wideband eigenmode may be shut off (i.e., not used) and the transmit power for this wideband eigenmode may be redistributed to other wideband eigenmodes.

[1105] The selected transmission mode q_m for each active wideband eigenmode is associated with a required SNR, $\gamma_{\text{req},m}$, that is equal to or lower than the operating SNR, i.e., $\gamma_{\text{req},m} \leq \gamma_{\text{op},m}$. The minimum transmit power required for each active wideband eigenmode, $P_{\text{req},m}$, may be computed as:

$$P_{\text{req},m} = \frac{P_m \cdot \gamma_{\text{req},m}}{\gamma_{\text{op},m}} \quad , \text{ for } m \in M . \qquad \text{Eq (22)}$$

The required transmit power is equal to zero ($P_{\text{req},m} = 0$) for each wideband eigenmode that is shut off (i.e., with transmission mode having index 0 in Table 2).

[1106] The excess power for each wideband eigenmode, $P_{\text{excess},m}$, is the amount of allocated power that is over the minimum power needed to achieve the required SNR (i.e., $P_{\text{excess},m} = P_m - P_{\text{req},m}$). The total excess power for all wideband eigenmodes, P_{excess}, may be computed as:

$$P_{\text{excess}} = \sum_{m=1}^{N_S} (P_m - P_{\text{req},m}) \quad . \qquad \text{Eq (23)}$$

[1107] The total excess power, P_{excess}, may be redistributed in various manners. For example, the total excess power, P_{excess}, may be redistributed to one or more wideband eigenmodes such that higher overall throughput is achieved. In one embodiment, the total excess power, P_{excess}, is redistributed to one unsaturated wideband eigenmode at a time, starting with the best one having the highest data rate, to move the wideband eigenmode to the next higher data rate. In another embodiment, the total excess power, P_{excess}, is redistributed to the wideband eigenmode that can achieve the highest increase in data rate with the least amount of transmit power.

[1108] If all wideband eigenmodes are operated at the highest data rate, or if the remaining excess power cannot increase the data rate of any wideband eigenmode, then the remaining excess power may be redistributed to one, multiple, or all active wideband eigenmodes to improve the SNR margins for these wideband eigenmodes.

E. Transmission Mode Adjustment

[1109] For step 822 in FIG. 8, the transmission mode for each wideband eigenmode may be adjusted based on information from the outer loop. The selected transmission modes for the downlink and uplink wideband eigenmodes may be adjusted using the techniques described above for FIG. 2. For example, if excessive packet errors are received on a given wideband eigenmode, then the outer loop may provide a transmission mode adjustment for that wideband eigenmode. As another example, a running average of the received SNRs may be maintained for each wideband eigenmode and used to compute the SNR margin for that wideband eigenmode. If the SNR margin for a given wideband eigenmode is negative, then the transmission mode for the wideband eigenmode may be adjusted to the next lower data rate. If a packet is transmitted across multiple wideband eigenmodes, then the transmission mode for the wideband eigenmode with the worse SNR margin may be adjusted to the next lower data rate whenever packet errors are detected. In any case, a transmission mode adjustment may direct the selection of another transmission mode with a lower data rate than the one selected in step 818.

II. MIMO-OFDM System

[1110] **FIG. 9A** shows a block diagram of an embodiment of an access point 510x and a user terminal 520x in the exemplary TDD MIMO-OFDM system. Access point 510x is one of access points 510 in FIG. 5, and user terminal 520x is one of user terminals 520. FIG. 9A shows the processing for downlink transmission. In this case, access point 510x is transmitter 110 in FIG. 1 and user terminal 520x is receiver 150.

[1111] For downlink transmission, at access point 510x, traffic data is provided from a data source 912 to a TX data processor 920, which demultiplexes the traffic data into N_C data streams, where $N_C > 1$. Traffic data may come from multiple data sources (e.g., one data source for each higher layer application) and the demultiplexing may not be needed. For simplicity, only one data source 912 is shown in FIG. 9A. TX data processor 920 formats, codes, interleaves, modulates, and scales each data stream in accordance with the transmission mode selected for that data stream to provide a corresponding scaled modulation symbol stream. The data rate, coding, and modulation for each data stream may be determined by a data rate control, a coding control, and a

modulation control, respectively, provided by a controller 940. TX data processor 920 provides N_C scaled modulation symbol streams to a TX spatial processor 928.

[1112] TX spatial processor 928 processes the N_C scaled modulation symbol streams based on a selected transmission scheme, multiplexes in pilot symbols, and provides N_{ap} transmit symbol streams to N_{ap} transmitter units (TMTR) 930a through 930ap. The selected transmission scheme may be for transmit diversity, spatial multiplexing, or beam-steering. Transmit diversity entails transmitting data redundantly from multiple antennas and/or on multiple subbands to obtain diversity and improve reliability. A space-time transmit diversity (STTD) may be used for transmit diversity. Beam-steering entails transmitting data on a single (best) spatial channel at full power using the phase steering information for the principal eigenmode. Spatial multiplexing entails transmitting data on multiple spatial channels to achieve higher spectral efficiency. The spatial processing for spatial multiplexing is shown in Table 1. Each transmitter unit 930 performs OFDM processing on its transmit symbol stream to provide a corresponding OFDM symbol stream, which is further processed to generate a modulated signal. The N_{ap} modulated signals from transmitter units 930a through 930ap are then transmitted via N_{ap} antennas 932a through 932ap, respectively.

[1113] At user terminal 520x, the N_{ap} transmitted signals are received by each of N_{ut} antennas 952a through 952ut, and the received signal from each antenna is provided to an associated receiver unit (RCVR) 954. Each receiver unit 954 conditions and digitizes its received signal to provide a stream of samples, which is further processed to provide a corresponding stream of received symbols. Receiver units 954a through 954ut provide N_{ut} received symbol streams to an RX spatial processor 962, which performs spatial processing based on the selected transmission scheme (e.g., as shown in Table 1 for spatial multiplexing). RX spatial processor 962 provides N_C recovered symbol streams, which are estimates of the N_C modulation symbol streams transmitted by access point 510x. An RX data processor 964 then demodulates, deinterleaves, and decodes each recovered symbol stream in accordance with the selected transmission mode to provide corresponding decoded data streams, which are estimates of the data streams transmitted by access point 510x. The processing by RX spatial processor 962 and RX data processor 964 is complementary to that performed by TX spatial processor 928 and TX data processor 920, respectively, at access point 510x.

[1114] A channel estimator 974 obtains estimates of one or more channel characteristics of the downlink and provides channel estimates to a controller 970. The channel estimates may be for channel gains, noise floor $N_{0,\text{ut}}$, and so on. RX data processor 964 may provide the status of each received data packet. Based on the various types of information received from channel estimator 974 and RX data processor 964, controller 970 determines a transmission mode for each of the multiple parallel channels on the downlink using the techniques described above. Each parallel channel may correspond to a wideband eigenmode (as described above) or some other combination of subbands and eigenmodes. Controller 970 provides feedback information, which may include the N_C selected transmission modes for the downlink, the channel estimates, the terminal noise floor, ACKs and/or NAKs for the receive data packets, and so on, or any combination thereof. The feedback information is processed by a TX data processor 978 and a TX spatial processor 980, multiplexed with a steered reference, conditioned by transmitter units 954a through 954ut, and transmitted via antennas 952a through 952ut to access point 510x.

[1115] At access point 510x, the N_{ut} transmitted signals from user terminal 520x are received by antennas 932a through 932ap, conditioned by receiver units 930a through 930ap, and processed by an RX spatial processor 934 and an RX data processor 936 to recover the feedback information sent by user terminal 520x. The feedback information is then provided to controller 940 and used to control the processing of the N_C data streams sent to user terminal 520x. For example, the data rate, coding, and modulation of each downlink data stream may be determined based on the transmission mode selected by user terminal 520x. The received ACK/NAK may be used to initiate either a full retransmission or an incremental transmission of each data packet received in error by user terminal 520x. For an incremental transmission, a small portion of a data packet received in error is transmitted to allow user terminal 520x to recover the packet.

[1116] A channel estimator 944 obtains channel gain estimates based on the received steered reference. The channel gain estimates are provided to controller 940 and used (possibly along with the user terminal noise floor $N_{0,\text{ut}}$ estimate) to derive transmit weights for the downlink. Controller 940 provides the data rate controls to data source 912 and TX data processor 920. Controller 940 further provides the coding and modulation controls and the transmit weights to TX data processor 920. The channel

estimation and transmission mode selection for downlink transmission may be performed as described above.

[1117] Controllers 940 and 970 direct the operation at access point 510x and user terminal 520x, respectively. Memory units 942 and 972 provide storage for program codes and data used by controllers 940 and 970, respectively.

[1118] **FIG. 9B** shows access point 510x and user terminal 520x for uplink transmission. In this case, user terminal 520x is transmitter 110 in FIG. 1 and access point 510x is receiver 150. The channel estimation and transmission mode selection for uplink transmission may be performed as described above. The data processing at access point 510x and user terminal 520x for uplink transmission may be performed in a manner similar to that described above for downlink transmission. The spatial processing at access point 510x and user terminal 520x for uplink transmission may be performed as shown in Table 1.

A. <u>Transmitter and Receiver Subsystems</u>

[1119] For clarity, the processing at access point 510x and user terminal 520x for downlink transmission is described in further detail below.

[1120] **FIG. 10** shows a block diagram of a transmitter subsystem 1000, which is an embodiment of the transmitter portion of access point 510x. For this embodiment, TX data processor 920 includes a demultiplexer (Demux) 1010, N_C encoders 1012a through 1012s, N_C channel interleavers 1014a through 1014s, N_C symbol mapping units 1016a through 1016s, and N_C signal scaling units 1018a through 1018s (i.e., one set of encoder, channel interleaver, symbol mapping unit, and signal scaling unit for each of the N_C data streams). Demultiplexer 1010 demultiplexes the traffic data (i.e., the information bits) into N_C data streams, where each data stream is provided at the data rate indicated by the data rate control. Demultiplexer 1010 may be omitted if traffic data is already provided as N_C data streams.

[1121] Each encoder 1012 receives and codes a respective data stream based on the selected coding scheme (as indicated by the coding control) to provide code bits. Each data stream may carry one or more data packets, and each data packet is typically coded separately to obtain a coded data packet. The coding increases the reliability of the data transmission. The selected coding scheme may include any combination of CRC coding, convolutional coding, turbo coding, block coding, and so on. The code bits from each encoder 1012 are provided to a respective channel interleaver 1014, which

interleaves the code bits based on a particular interleaving scheme. If the interleaving is dependent on transmission mode, then controller 940 provides an interleaving control (as indicated by the dashed line) to channel interleaver 1014. The interleaving provides time, frequency, and/or spatial diversity for the code bits.

[1122] The interleaved bits from each channel interleaver 1014 are provided to a respective symbol mapping unit 1016, which maps the interleaved bits based on the selected modulation scheme (as indicated by the modulation control) to provide modulation symbols. Unit 1016 groups each set of B interleaved bits to form a B-bit binary value, where $B \geq 1$, and further maps each B-bit value to a specific modulation symbol based on the selected modulation scheme (e.g., QPSK, M-PSK, or M-QAM, where $M = 2^B$). Each modulation symbol is a complex value in a signal constellation defined by the selected modulation scheme. The modulation symbols from each symbol mapping unit 1016 are then provided to a respective signal scaling unit 1018, which scales the modulation symbols with the transmit weights, $W_m(k)$ for $k \in K$, to achieve channel inversion and power distribution. Signal scaling units 1018a through 1018s provide N_C scaled modulation symbol streams.

[1123] Each data stream is transmitted on a respective parallel channel that may include any number and any combination of subbands, transmit antennas, and spatial channels. For example, one data stream may be transmitted on all usable subbands of each wideband eigenmode, as described above. TX spatial processor 928 performs the required spatial processing, if any, on the N_C scaled modulation symbol streams and provides N_{ap} transmit symbol streams. The spatial processing may be performed as shown in Table 1.

[1124] For a transmission scheme whereby one data stream is transmitted on all subbands of each wideband eigenmode (for a full-CSI MIMO system, as described above), N_S sets of encoder 1012, channel interleaver 1014, symbol mapping unit 1016, and signal scaling unit 1018 may be used to process N_S data streams (where $N_C = N_S = N_{ap} \leq N_{ut}$ for a full rank channel response matrix) to provide N_{ap} scaled modulation symbol streams. TX spatial processor 928 then performs spatial processing on the N_{ap} scaled modulation symbol streams, as shown in Table 1, to provide the N_{ap} transmit symbol streams.

[1125] For a transmission scheme whereby one data stream is transmitted on all subbands of each transmit antenna (for a partial-CSI MIMO system), N_{ap} sets of encoder

1012, channel interleaver 1014, symbol mapping unit 1016, and signal scaling unit 1018 may be used to process N_{ap} data streams (where $N_C = N_{ap}$) to provide N_{ap} scaled modulation symbol streams. TX spatial processor 928 then simply passes each scaled modulation symbol stream as a transmit symbol stream. Since spatial processing is not performed for this transmission scheme, each transmit symbol is a modulation symbol.

[1126] In general, TX spatial processor 928 performs the appropriate demultiplexing and/or spatial processing of the scaled modulation symbols to obtain transmit symbols for the parallel channel used for each data stream. TX spatial processor 928 further multiplexes pilot symbols with the transmit symbols, e.g., using time division multiplex (TDM) or code division multiplex (CDM). The pilot symbols may be sent in all or a subset of the subbands/eigenmodes used to transmit traffic data. TX spatial processor 928 provides N_{ap} transmit symbol streams to N_{ap} transmitter units 930a through 930ap.

[1127] Each transmitter unit 930 performs OFDM processing on a respective transmit symbol stream and provides a corresponding modulated signal. The OFDM processing typically includes (1) transforming each set of N_F transmit symbols to the time domain using an N_F-point inverse fast Fourier transform (IFFT) to obtain a "transformed" symbol that contains N_F samples and (2) repeating a portion of each transformed symbol to obtain an OFDM symbol that contains $N_F + N_{cp}$ samples. The repeated portion is referred to as the cyclic prefix, and N_{cp} indicates the number of samples being repeated. The OFDM symbols are further processed (e.g., converted to one or more analog signals, amplified, filtered, and frequency upconverted) by transmitter unit 930 to generate the modulated signal. Other designs for transmitter subsystem 1000 may also be implemented and are within the scope of the invention.

[1128] Controller 940 may perform various functions related to closed-loop rate control for the downlink and uplink (e.g., transmission mode selection for the uplink and transmit weight computation for the downlink). For uplink transmission, controller 940 may perform process 800 in FIG. 8 and selects a transmission mode for each of the multiple parallel channels on the uplink. Within controller 940, a power allocation unit 1042 distributes the total transmit power, $P_{\text{total,up}}$, to the multiple parallel channels (e.g., based on the channel gain estimates $\hat{\hat{\sigma}}_m(k)$ and the noise floor estimate $N_{0,\text{ap}}$ for the access point). A channel inversion unit 1044 performs channel inversion for each

parallel channel. A transmission mode (TM) selector 1046 selects a suitable transmission mode for each parallel channel. Memory unit 942 may store a look-up table 1048 for supported transmission modes and their required SNRs (e.g., as shown in Table 2). For downlink transmission, controller 940 may also perform process 800 in FIG. 8 to determine the transmit power for each subband of each wideband eigenmode and computes the transmit weights used for scaling modulation symbols prior to transmission on the downlink.

[1129] **FIG. 11** shows a block diagram of a receiver subsystem 1100, which is an embodiment of the receiver portion of user terminal 520x. The N_{ap} transmitted signals from access point 510x are received by antennas 952a through 952ut, and the received signal from each antenna is provided to a respective receiver unit 954. Each receiver unit 954 conditions and digitizes its received signal to obtain a stream of samples, and further performs OFDM processing on the samples. The OFDM processing at the receiver typically includes (1) removing the cyclic prefix in each received OFDM symbol to obtain a received transformed symbol and (2) transforming each received transformed symbol to the frequency domain using a fast Fourier transform (FFT) to obtain a set of N_F received symbols for the N_F subbands. The received symbols are estimates of the transmit symbols sent by access point 510x. Receiver units 954a through 954ut provide N_{ut} received symbol streams to RX spatial processor 962.

[1130] RX spatial processor 962 performs spatial or space-time processing on the N_{ut} received symbol streams to provide N_C recovered symbol streams. RX spatial processor 962 may implement a linear zero-forcing (ZF) equalizer (which is also referred to as a channel correlation matrix inversion (CCMI) equalizer), a minimum mean square error (MMSE) equalizer, an MMSE linear equalizer (MMSE-LE), a decision feedback equalizer (DFE), or some other equalizer.

[1131] RX data processor 964 receives the N_C recovered symbol streams from RX spatial processor 962. Each recovered symbol stream is provided to a respective symbol demapping unit 1132, which demodulates the recovered symbols in accordance with the modulation scheme used for that stream, as indicated by a demodulation control provided by controller 970. The demodulated data stream from each symbol demapping unit 1132 is de-interleaved by an associated channel de-interleaver 1134 in a manner complementary to that performed at access point 510x for that data stream. If the interleaving is dependent on transmission mode, then controller 970 provides a

deinterleaving control to channel de-interleaver 1134, as indicated by the dashed line. The de-interleaved data from each channel de-interleaver 1134 is decoded by an associated decoder 1136 in a manner complementary to that performed at access point510x, as indicated by a decoding control provided by controller 970. For example, a turbo decoder or a Viterbi decoder may be used for decoder 1136 if turbo or convolutional coding, respectively, is performed at access point 510x. Decoder 1136 may also provide the status of each received data packet (e.g., indicating whether the packet was received correctly or in error). Decoder 1136 may further store demodulated data for packets decoded in error, so that this data may be combined with additional data from a subsequent incremental transmission and decoded.

[1132] In the embodiment shown in FIG. 11, channel estimator 974 estimates the channel response and the noise floor at user terminal 520x (e.g., based on the received pilot symbols) and provides the channel estimates to controller 970. Controller 970 performs various functions related to closed-loop rate control for both the downlink and uplink (e.g., transmission mode selection for the downlink and transmit weight computation for the uplink). For downlink transmission, controller 970 may perform process 800 in FIG. 8. Within controller 970, a power allocation unit 1172 distributes the total transmit power, $P_{\text{total, dn}}$, to the multiple parallel channels (e.g., based on the channel gain estimates $\hat{\sigma}_m(k)$ and the noise floor $N_{0,\text{ut}}$ estimate for the user terminal). A channel inversion unit 1174 performs channel inversion for each of the multiple parallel channels. A transmission mode (TM) selector 1176 selects a suitable transmission mode for each parallel channel. Memory unit 972 may store a look-up table 1178 for supported transmission modes and their required SNRs (e.g., as shown in Table 2). Controller 970 provides N_C selected transmission modes for the N_C parallel channels on the downlink, which may be part of the feedback information sent to access point 510x. For uplink transmission, controller 970 may also perform process 800 in FIG. 8 to determine the transmit power for each subband of each wideband eigenmode and computes the transmit weights used for scaling modulation symbols prior to transmission on the uplink.

[1133] For clarity, transmitter subsystem 1000 has been described for access point 510x and receiver subsystem 1100 has been described for user terminal 520x. Transmitter subsystem 1000 may also be used for the transmitter portion of user

terminal 520x, and receiver subsystem 1100 may also be used for the receiver portion of access point 510x.

B. Downlink and Uplink Rate Control

[1134] **FIG. 12A** shows a process for performing closed-loop rate control for the downlink based on the frame structure shown in FIG. 6. A BCH PDU is transmitted in the first segment of each TDD frame (see FIG. 6) and includes the MIMO pilot that can be used by the user terminal to estimate and track the downlink. A steered reference may also be sent in the preamble of an FCH PDU sent to the user terminal. The user terminal estimates the downlink based on the MIMO pilot and/or the steered reference and selects a suitable transmission mode (with the highest supported data rate) for each downlink wideband eigenmode (i.e., each parallel channel). The user terminal then sends these transmission modes as "proposed" transmission modes for the downlink in an RCH PDU sent to the access point.

[1135] The access point receives the proposed transmission modes from the user terminal and schedules data transmission on the downlink in subsequent TDD frame(s). The access point selects the transmission modes for the downlink, which may be the ones received from the user terminal or some other transmission modes (with lower data rates), depending on system loading and other factors. The access point sends assignment information for the user terminal (which includes the transmission modes selected by the access point for downlink transmission) on the FCCH. The access point then transmits data on the FCH to the user terminal using the selected transmission modes. The user terminal receives the assignment information and obtains the transmission modes selected by the access point. The user terminal then processes the downlink transmission in accordance with the selected transmission mode. For the embodiment shown in FIG. 12A, the delay between the channel estimation and transmission mode selection by the user terminal and the use of these transmission modes for downlink transmission is typically one TDD frame, but may be different depending on applications, system configurations, and other factors.

[1136] **FIG. 12B** shows a process for performing closed-loop rate control for the uplink based on the frame structure shown in FIG. 6. The user terminal transmits a steered reference on the RACH during system access and on the RCH upon being assigned FCH/RCH resources (see FIG. 6). The access point estimates the uplink based on the received steered reference and selects a suitable transmission mode for each

uplink wideband eigenmode. The access point sends assignment information for the user terminal (which includes the transmission modes selected for uplink transmission) on the FCCH. The user terminal transmits data on the RCH to the access point using the selected transmission modes. The access point processes the uplink transmission in accordance with the selected transmission modes.

[1137] The closed-loop rate control techniques described herein may be implemented by various means. For example, these techniques may be implemented in hardware, software, or a combination thereof. For a hardware implementation, the elements used for closed-loop rate control at the transmitter and the receiver (e.g., controllers 940 and 970) may be implemented within one or more application specific integrated circuits (ASICs), digital signal processors (DSPs), digital signal processing devices (DSPDs), programmable logic devices (PLDs), field programmable gate arrays (FPGAs), processors, controllers, micro-controllers, microprocessors, other electronic units designed to perform the functions described herein, or a combination thereof.

[1138] For a software implementation, portions of the closed-loop rate control may be implemented with modules (e.g., procedures, functions, and so on) that perform the functions described herein. The software codes may be stored in a memory unit (e.g., memory unit 942 or 972 in FIGS. 9A and 9B) and executed by a processor (e.g., controller 940 or 970). The memory unit may be implemented within the processor or external to the processor, in which case it can be communicatively coupled to the processor via various means as is known in the art.

[1139] Headings are included herein for reference and to aid in locating certain sections. These headings are not intended to limit the scope of the concepts described therein under, and these concepts may have applicability in other sections throughout the entire specification.

[1140] The previous description of the disclosed embodiments is provided to enable any person skilled in the art to make or use the present invention. Various modifications to these embodiments will be readily apparent to those skilled in the art, and the generic principles defined herein may be applied to other embodiments without departing from the spirit or scope of the invention. Thus, the present invention is not intended to be limited to the embodiments shown herein but is to be accorded the widest scope consistent with the principles and novel features disclosed herein.

WHAT IS CLAIMED IS:

CLAIMS

1. A method of transmitting data on a plurality of parallel channels in a wireless communication system, comprising:

obtaining channel estimates for each of the plurality of parallel channels;

selecting a transmission mode for each of the plurality of parallel channels based on the channel estimates for the parallel channel, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel; and

sending the transmission mode for each of the plurality of parallel channels to a transmitting entity, wherein a data transmission on each of the plurality of parallel channels is processed at the transmitting entity in accordance with the transmission mode selected for the parallel channel.

2. The method of claim 1, further comprising:

receiving data transmissions on the plurality of parallel channels from the transmitting entity; and

processing the data transmissions in accordance with the transmission mode selected for each of the plurality of parallel channels to recover data sent on the parallel channel.

3. The method of claim 1, wherein the channel estimates for each of the plurality of parallel channels include at least one channel gain estimate and a noise floor estimate for the parallel channel.

4. The method of claim 1, wherein the selecting includes

determining a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel, and wherein the transmission mode for each of the plurality of parallel channels is selected based on the received SNR for the parallel channel.

5. The method of claim 4, wherein the selecting further includes

determining an SNR offset for each of the plurality of parallel channels, and wherein the transmission mode for each of the plurality of parallel channels is further selected based on the SNR offset for the parallel channel.

6. The method of claim 5, wherein the selecting further includes

determining an operating SNR for each of the plurality of parallel channels based on the received SNR and the SNR offset for the parallel channel, and wherein the transmission mode for each of the plurality of parallel channels is selected based on the operating SNR for the parallel channel.

7. The method of claim 6, wherein the transmission mode for each of the plurality of parallel channels is further selected based on a set of required SNRs for a set of transmission modes supported by the system.

8. The method of claim 1, further comprising:

estimating the quality of the data transmission received on each of the plurality of parallel channels, and wherein the transmission mode for each of the plurality of parallel channels is further selected based on the estimated quality of the data transmission received on the parallel channel.

9. The method of claim 5, further comprising:

adjusting the SNR offset for each of the plurality of parallel channels based on status of data packets received on the parallel channel.

10. The method of claim 5, further comprising:

adjusting the SNR offset for each of the plurality of parallel channels based on at least one decoder metric maintained for the parallel channel.

11. The method of claim 1, further comprising:

detecting for packet errors for each of the plurality of parallel channels; and

adjusting the transmission mode for each of the plurality of parallel channels based on the packet errors for the parallel channel.

12. The method of claim 4, further comprising:

determining an SNR margin for each of the plurality of parallel channels based on the received SNR and a required SNR for the parallel channel; and

adjusting the transmission mode for each of the plurality of parallel channels based on SNR margins for the plurality of parallel channels.

13. The method of claim 6, further comprising:

distributing total transmit power to the plurality of parallel channels, and wherein the operating SNR for each of the plurality of parallel channels is further determined based on transmit power distributed to the parallel channel.

14. The method of claim 13, wherein the total transmit power is uniformly distributed to the plurality of parallel channels.

15. The method of claim 13, wherein the total transmit power is distributed to the plurality of parallel channels using a water-filling procedure.

16. The method of claim 13, further comprising:

determining excess power for each of the plurality of parallel channels based on the operating SNR for the parallel channel, a required SNR for the transmission mode selected for the parallel channel, and the transmit power distributed to the parallel channel;

accumulating the excess power for each of the plurality of parallel channels to obtain total excess power for the plurality of parallel channels; and

redistributing the total excess power to at least one of the plurality of parallel channels.

17. The method of claim 16, wherein the total excess power is redistributed evenly to unsaturated parallel channels among the plurality of parallel channels, where the unsaturated parallel channels have data rates greater than zero and less than a maximum data rate.

18. The method of claim 16, wherein the total excess power is redistributed to one parallel channel, selected from among the plurality of parallel channels, that can achieve a highest increase in data rate with the total excess power.

19. The method of claim 13, wherein each of the plurality of parallel channels includes a plurality of subbands, the method further comprising:

distributing the transmit power for each of the plurality of parallel channels across the plurality of subbands of the parallel channel to achieve similar received SNRs for the plurality of subbands.

20. The method of claim 13, wherein each of the plurality of parallel channels includes a plurality of subbands, the method further comprising:

distributing the transmit power for each of the plurality of parallel channels uniformly across the plurality of subbands of the parallel channel.

21. The method of claim 1, wherein the wireless communication system is an orthogonal frequency division multiplex (OFDM) communication system, and wherein the plurality of parallel channels are formed by a plurality of disjoint sets of subbands.

22. The method of claim 1, wherein the wireless communication system is a frequency division multiplex (FDM) communication system, and wherein the plurality of parallel channels are formed by a plurality of frequency subbands.

23. The method of claim 1, wherein the wireless communication system is a time division multiplex (TDM) communication system, and wherein the plurality of parallel channels are formed by a plurality of time slots.

24. The method of claim 1, wherein the wireless communication system is a multiple-input multiple-output (MIMO) communication system, and wherein the plurality of parallel channels are formed by a plurality of spatial channels.

25. The method of claim 1, wherein the wireless communication system is a multiple-input multiple-output (MIMO) communication system with orthogonal frequency division multiplex (OFDM).

26. The method of claim 25, wherein the plurality of parallel channels are formed by a plurality of wideband spatial channels, and wherein each of the plurality of parallel channels includes a plurality of subbands.

27. The method of claim 25, wherein the channel estimates for each of the plurality of parallel channels are obtained based on a pilot transmitted from each of a plurality of antennas by the transmitting entity.

28. The method of claim 25, wherein the channel estimates for each of the plurality of parallel channels are obtained based on a steered reference transmitted from a plurality of antennas by the transmitting entity.

29. An apparatus in a wireless communication system, comprising:

means for obtaining channel estimates for each of a plurality of parallel channels;

means for selecting a transmission mode for each of the plurality of parallel channels based on the channel estimates for the parallel channel, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel; and

means for sending the transmission mode for each of the plurality of parallel channels to a transmitting entity, wherein a data transmission on each of the plurality of parallel channels is processed at the transmitting entity in accordance with the transmission mode selected for the parallel channel.

30. The apparatus of claim 29, further comprising:

means for receiving data transmissions on the plurality of parallel channels from the transmitting entity; and

means for processing the received data transmissions in accordance with the transmission mode selected for each of the plurality of parallel channels to recover data sent on the parallel channel.

31. The apparatus of claim 29, wherein the means for selecting includes

means for determining a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel, and wherein the transmission mode for each of the plurality of parallel channels is selected based on the received SNR for the parallel channel.

32. The apparatus of claim 29, further comprising:

means for estimating the quality of the data transmission received on each of the plurality of parallel channels, and wherein the transmission mode for each of the plurality of parallel channels is further selected based on the estimated quality of the data transmission received on the parallel channel.

33. An apparatus in a wireless communication system, comprising:

a channel estimator operative to obtain channel estimates for each of a plurality of parallel channels; and

a controller operative to select a transmission mode for each of the plurality of parallel channels based on the channel estimates for the parallel channel, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel, and wherein a data transmission on each of the plurality of parallel channels is processed at a transmitting entity in accordance with the transmission mode selected for the parallel channel.

34. The apparatus of claim 33, further comprising:

a receive (RX) data processor operative to receive data transmissions on the plurality of parallel channels and to process the received data transmissions in accordance with the transmission mode selected for each of the plurality of parallel channels to recover data sent on the parallel channel.

35. The apparatus of claim 33, wherein the controller is operative to determine a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel and to select the transmission mode for each parallel channel based on the received SNR for the parallel channel.

36. The apparatus of claim 33, wherein the controller is operative to obtain an estimate of the quality of the data transmission received on each of the plurality of parallel channels and to adjust the transmission mode for each parallel channel based on the estimated quality of the data transmission received on the parallel channel.

37. A method of transmitting data on a plurality of parallel channels in a wireless communication system, comprising:

receiving feedback information from a receiving entity, wherein the feedback information is indicative of the quality of the plurality of parallel channels;

determining a transmission mode for each of the plurality of parallel channels based on the feedback information, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel;

processing data for each of the plurality of parallel channels in accordance with the transmission mode for the parallel channel; and

transmitting the processed data for each of the plurality of parallel channels on the parallel channel to the receiving entity.

38. The method of claim 37, wherein the transmission mode for each of the plurality of parallel channels is selected by the receiving entity based on channel estimates obtained for the parallel channel, and wherein the feedback information includes a plurality of transmission modes selected by the receiving entity for the plurality of parallel channels.

39. The method of claim 37, further comprising:

obtaining channel gain estimates for each of the plurality of parallel channels, and wherein the transmission mode for each of the plurality of parallel channels is determined based on the channel gain estimates for the parallel channel and a noise floor estimate for the parallel channel included in the feedback information from the receiving entity.

40. The method of claim 39, wherein the channel gain estimates for each of the plurality of parallel channels are obtained based on a steered reference received from the receiving entity.

41. The method of claim 37, further comprising:

receiving an adjustment to the transmission mode for a first parallel channel among the plurality of parallel channels; and

processing data for the first parallel channel in accordance with the adjustment to the transmission mode for the first parallel channel.

42. The method of claim 41, wherein the adjustment to the transmission mode for the first parallel channel is determined based on packet errors detected for the first parallel channel.

43. The method of claim 41, wherein the adjustment to the transmission mode for the first parallel channel is determined based on a received signal-to-noise ratio (SNR) and a required SNR for the first parallel channel.

44. The method of claim 37, further comprising:

computing, for each of the plurality of parallel channels, a plurality of transmit weights for a plurality of subbands of the parallel channel, wherein the plurality of transmit weights achieve similar received signal-to-noise ratios (SNRs) for the plurality of subbands of the parallel channel; and

scaling the processed data for each of the plurality of parallel channels with the plurality of transmit weights for the parallel channel, and wherein the scaled and processed data for each of the plurality of parallel channels is transmitted on the parallel channel.

45. An apparatus in a wireless communication system, comprising:

means for receiving feedback information from a receiving entity, wherein the feedback information is indicative of the quality of the plurality of parallel channels;

means for determining a transmission mode for each of a plurality of parallel channels based on the feedback information, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel;

means for processing data for each of the plurality of parallel channels in accordance with the transmission mode for the parallel channel; and

means for transmitting the processed data for each of the plurality of parallel channels on the parallel channel.

46. The apparatus of claim 45, further comprising:

means for obtaining channel gain estimates for each of the plurality of parallel channels, and wherein the transmission mode for each of the plurality of parallel channels is determined based on the channel gain estimates for the parallel channel and a noise floor estimate for the parallel channel included in the feedback information from the receiving entity.

47. The apparatus of claim 45, further comprising:

means for receiving an adjustment to the transmission mode for a first parallel channel among the plurality of parallel channels; and

means for processing data for the first parallel channel in accordance with the adjustment to the transmission mode for the first parallel channel

48. An apparatus in a wireless communication system, comprising:

a controller operative to determine a transmission mode for each of a plurality of parallel channels based on feedback information received from a receiving entity, wherein the feedback information is indicative of the quality of the plurality of parallel channels, and wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel;

a transmit (TX) data processor operative to process data for each of the plurality of parallel channels in accordance with the transmission mode for the parallel channel; and

at least one transmitter unit operative to transmit the processed data for each of the plurality of parallel channels on the parallel channel.

49. The apparatus of claim 48, wherein the controller if operative to obtain channel gain estimates for each of the plurality of parallel channels and to determine the transmission mode for each of the plurality of parallel channels based on the channel gain estimates for the parallel channel and a noise floor estimate for the parallel channel included in the feedback information from the receiving entity.

50. The apparatus of claim 48, wherein the controller is operative to obtain an adjustment to the transmission mode for a first parallel channel among the plurality of parallel channels, and wherein the TX data processor is operative to process data for the first parallel channel in accordance with the adjustment to the transmission mode for the first parallel channel.

51. A method of transmitting data on a plurality of parallel channels in a wireless communication system, comprising:

obtaining channel estimates for each of the plurality of parallel channels;

computing a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel;

computing an operating SNR for each of the plurality of parallel channels based on the received SNR and an SNR offset for the parallel channel;

selecting a transmission mode for each of the plurality of parallel channels based on the operating SNR for the parallel channel and a set of required SNRs for a set of transmission modes supported by the system, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel; and

processing data for each of the plurality of parallel channels in accordance with the transmission mode selected for the parallel channel.

52. The method of claim 51, further comprising:

estimating the quality of a data transmission received on each of the plurality of parallel channels; and

adjusting the SNR offset for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.

53. The method of claim 52, wherein the quality of the data transmission received on each of the plurality of parallel channels is estimated based on status of packets received on the parallel channel.

54. The method of claim 52, further comprising:

adjusting the transmission mode for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.

55. An apparatus in a wireless communication system, comprising:

means for obtaining channel estimates for each of a plurality of parallel channels;

means for computing a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel;

means for computing an operating SNR for each of the plurality of parallel channels based on the received SNR and an SNR offset for the parallel channel;

means for selecting a transmission mode for each of the plurality of parallel channels based on the operating SNR for the parallel channel and a set of required SNRs for a set of transmission modes supported by the system, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel; and

means for processing data for each of the plurality of parallel channels in accordance with the transmission mode selected for the parallel channel.

56. The apparatus of claim 55, further comprising:

means for estimating the quality of a data transmission received on each of the plurality of parallel channels; and

means for adjusting the SNR offset for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.

57. The method of claim 56, further comprising:

means for adjusting the transmission mode for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.

58. An apparatus in a wireless communication system, comprising:

a channel estimator operative to provide channel gain estimates for each of a plurality of parallel channels;

a selector operative to compute a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel, compute an operating SNR for each of the plurality of parallel channels based on the received SNR and an SNR offset for the parallel channel, and select a transmission mode for each of the plurality of parallel channels based on the operating SNR for the parallel channel and a set of required SNRs for a set of transmission modes supported by the system, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel; and

a data processor operative to process data for each of the plurality of parallel channels in accordance with the transmission mode selected for the parallel channel.

59. The apparatus of claim 58, wherein the selector is operative to receive an estimate of the quality of a data transmission received on each of the plurality of parallel channels and to adjust the SNR offset for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.

60. The method of claim 59, wherein the selector is further operative to adjust the transmission mode for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.

61. A processor readable media for storing instructions operable to:

obtain channel gain estimates for each of a plurality of parallel channels in a wireless communication system;

compute a received signal-to-noise ratio (SNR) for each of the plurality of parallel channels based on the channel estimates for the parallel channel;

compute an operating SNR for each of the plurality of parallel channels based on the received SNR and an SNR offset for the parallel channel; and

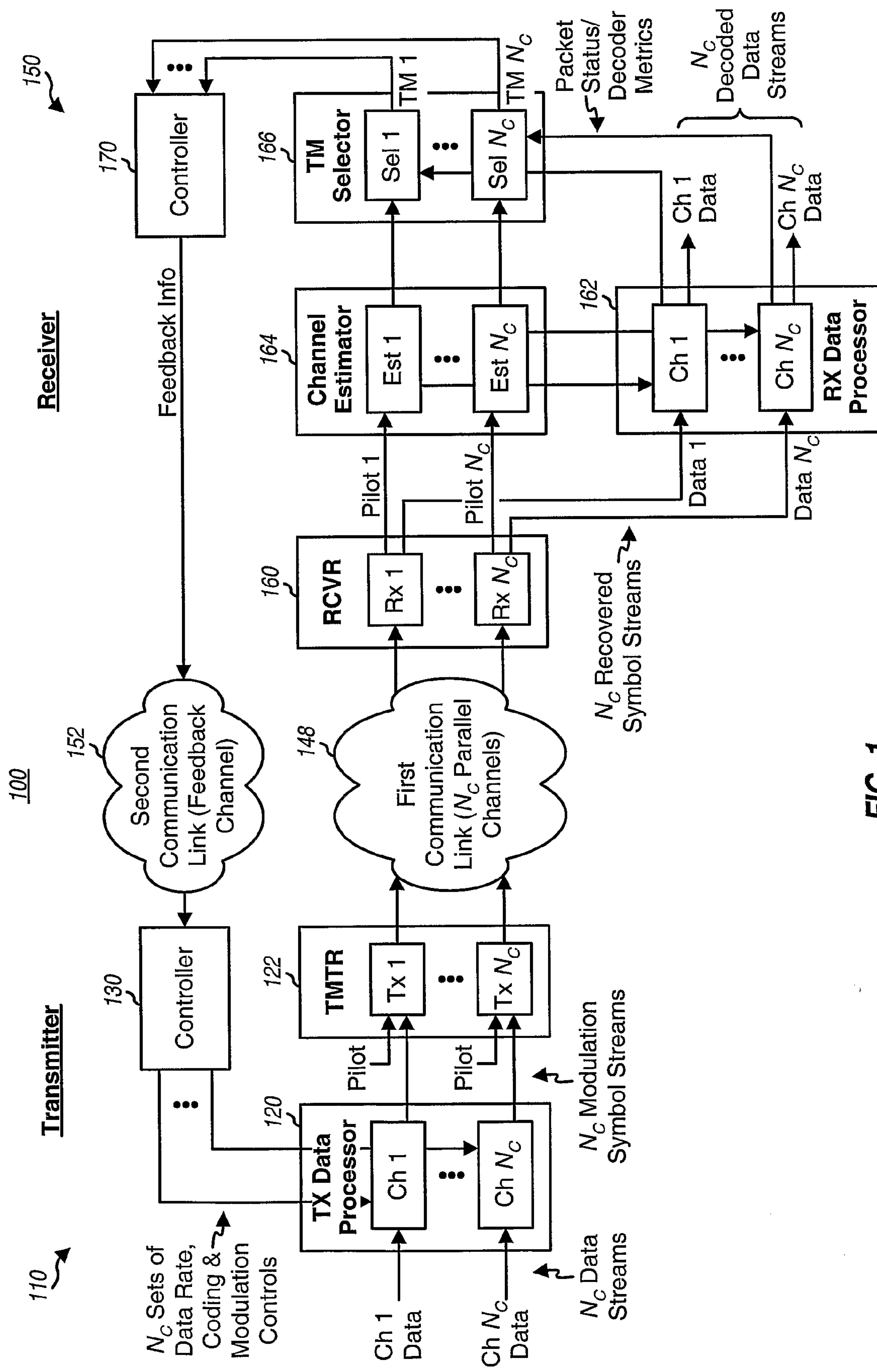
select a transmission mode for each of the plurality of parallel channels based on the operating SNR for the parallel channel and a set of required SNRs for a set of transmission modes supported by the system, wherein the transmission mode for each of the plurality of parallel channels indicates a data rate for the parallel channel, and wherein data is sent on each of the plurality of parallel channels in accordance with the transmission mode selected for the parallel channel.

62. The processor readable media of claim 61 and further storing instructions operable to:

adjust the SNR offset for each of the plurality of parallel channels based on an estimate of the quality of the data transmission received on the parallel channel.

63. The processor readable media of claim 62 and further storing instructions operable to:

adjust the transmission mode for each of the plurality of parallel channels based on the estimated quality of the data transmission received on the parallel channel.



100
110
Transmitter
Receiver
150
130
Controller
170
Controller
Feedback Info
152
Second Communication Link (Feedback Channel)
148
First Communication Link (N_C Parallel Channels)
120
TX Data Processor
Ch 1
Ch N_C
122
TMTR
Tx 1
Tx N_C
Pilot
Pilot
160
RCVR
Rx 1
Rx N_C
Pilot 1
Pilot N_C
Data 1
Data N_C
164
Channel Estimator
Est 1
Est N_C
162
RX Data Processor
Ch 1
Ch N_C
166
TM Selector
Sel 1
Sel N_C
TM 1
TM N_C
Packet Status/ Decoder Metrics
N_C Decoded Data Streams
Ch 1 Data
Ch N_C Data
Ch 1 Data
Ch N_C Data
N_C Data Streams
N_C Sets of Data Rate, Coding & Modulation Controls
N_C Modulation Symbol Streams
N_C Recovered Symbol Streams

FIG. 1

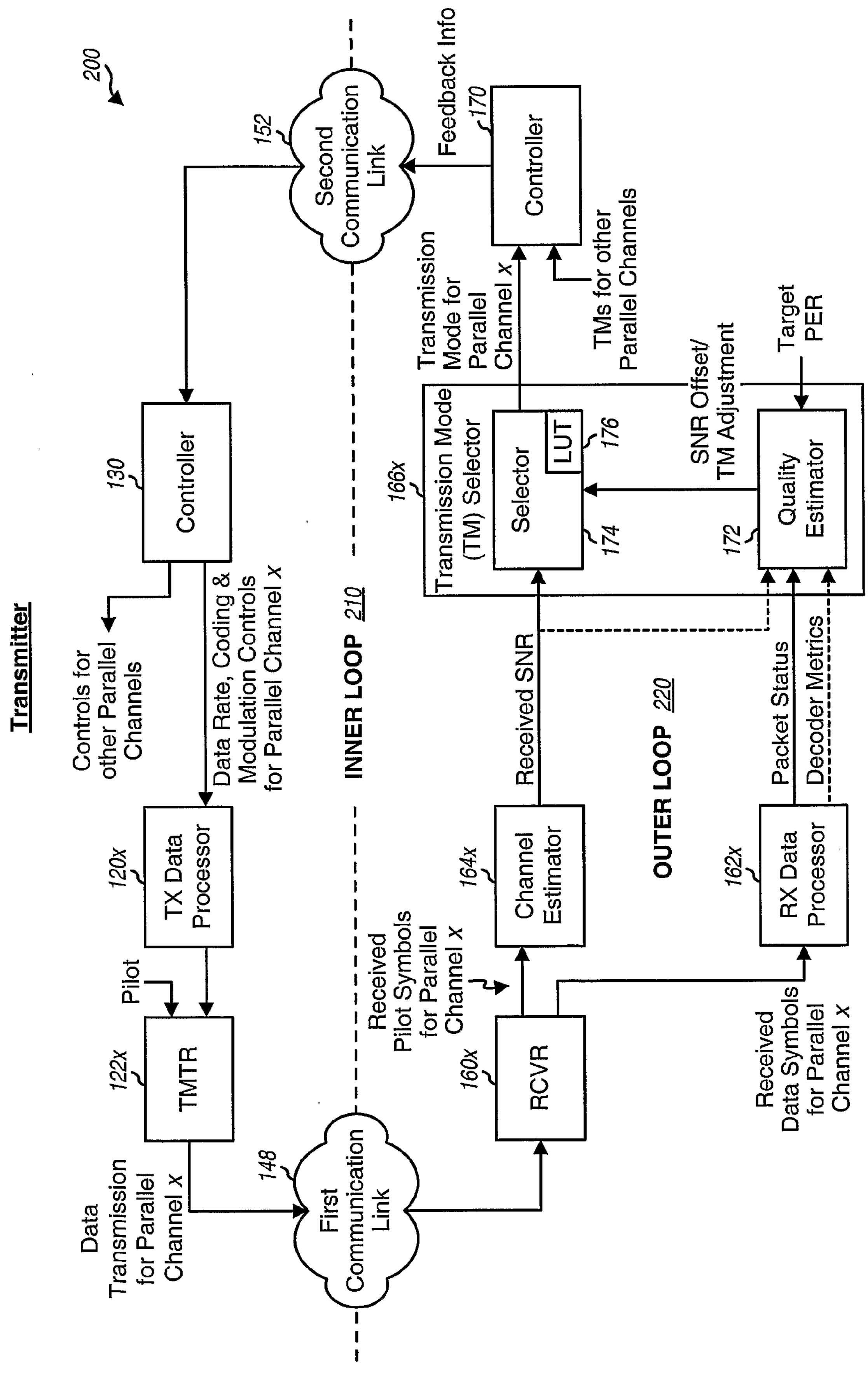
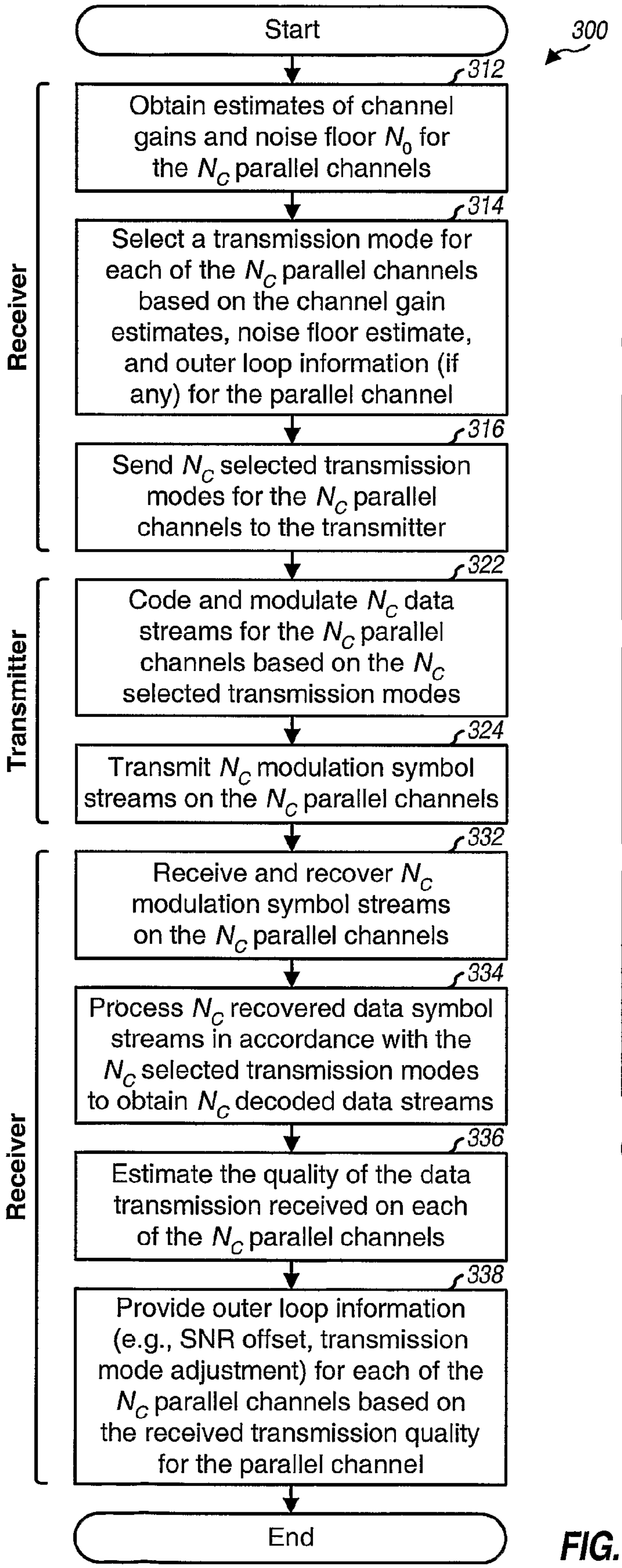
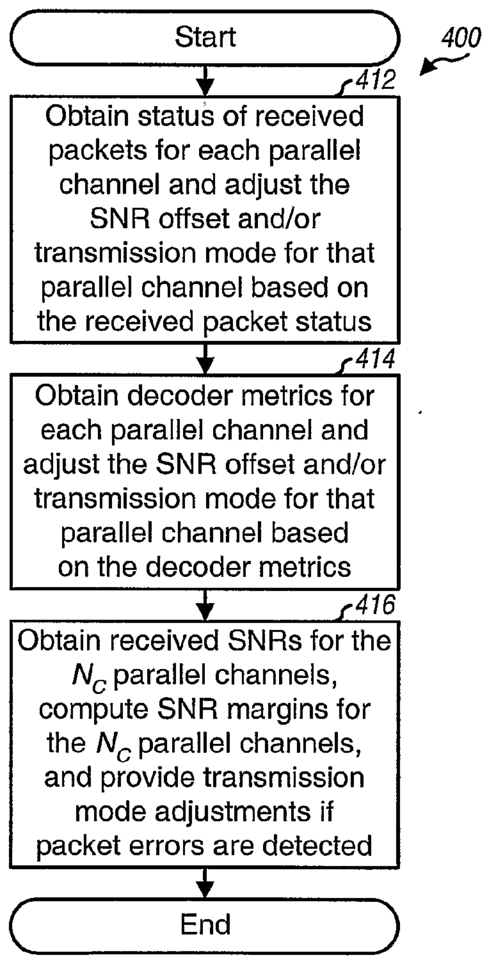

200
Transmitter
Receiver
Data Transmission for Parallel Channel x
122x
TMTR
Pilot
120x
TX Data Processor
130
Controller
Controls for other Parallel Channels
Data Rate, Coding & Modulation Controls for Parallel Channel x
148
First Communication Link
152
Second Communication Link
INNER LOOP 210
OUTER LOOP 220
160x
RCVR
Received Pilot Symbols for Parallel Channel x
164x
Channel Estimator
Received SNR
Received Data Symbols for Parallel Channel x
162x
RX Data Processor
Packet Status
Decoder Metrics
166x
Transmission Mode (TM) Selector
174
Selector
LUT
176
172
Quality Estimator
SNR Offset/ TM Adjustment
Target PER
Transmission Mode for Parallel Channel x
TMs for other Parallel Channels
170
Controller
Feedback Info

FIG. 2



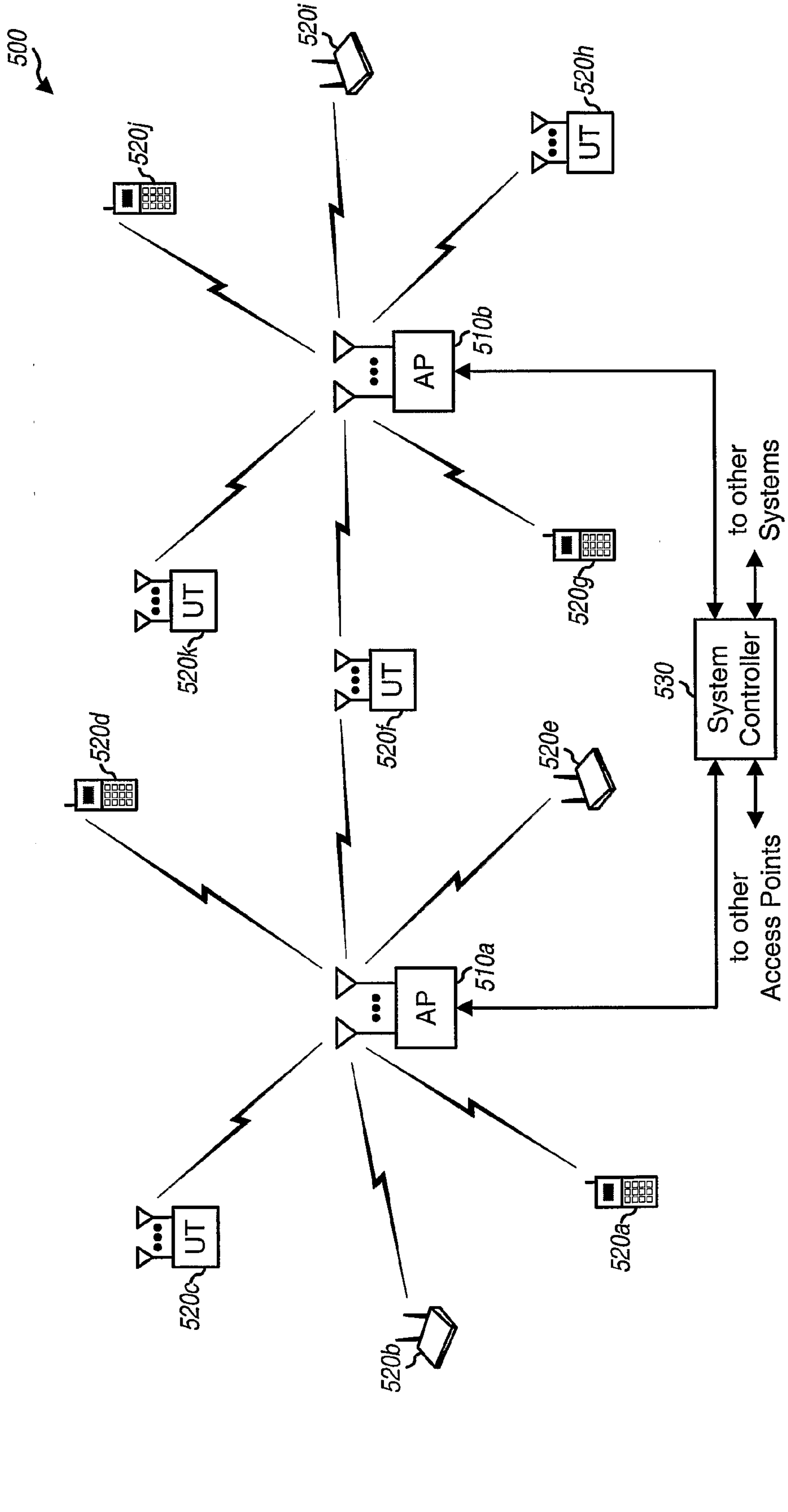
Start
300
312
Obtain estimates of channel gains and noise floor N_0 for the N_C parallel channels
314
Select a transmission mode for each of the N_C parallel channels based on the channel gain estimates, noise floor estimate, and outer loop information (if any) for the parallel channel
316
Send N_C selected transmission modes for the N_C parallel channels to the transmitter
Receiver
322
Code and modulate N_C data streams for the N_C parallel channels based on the N_C selected transmission modes
324
Transmit N_C modulation symbol streams on the N_C parallel channels
Transmitter
332
Receive and recover N_C modulation symbol streams on the N_C parallel channels
334
Process N_C recovered data symbol streams in accordance with the N_C selected transmission modes to obtain N_C decoded data streams
336
Estimate the quality of the data transmission received on each of the N_C parallel channels
338
Provide outer loop information (e.g., SNR offset, transmission mode adjustment) for each of the N_C parallel channels based on the received transmission quality for the parallel channel
Receiver
End

FIG. 3



Start
400
412
Obtain status of received packets for each parallel channel and adjust the SNR offset and/or transmission mode for that parallel channel based on the received packet status
414
Obtain decoder metrics for each parallel channel and adjust the SNR offset and/or transmission mode for that parallel channel based on the decoder metrics
416
Obtain received SNRs for the N_C parallel channels, compute SNR margins for the N_C parallel channels, and provide transmission mode adjustments if packet errors are detected
End

FIG. 4



500
520i
520h
UT
520j
AP
510b
to other Systems
520k
UT
UT
520f
520g
520d
520e
530
System Controller
AP
510a
to other Access Points
UT
520c
520b
520a

FIG. 5

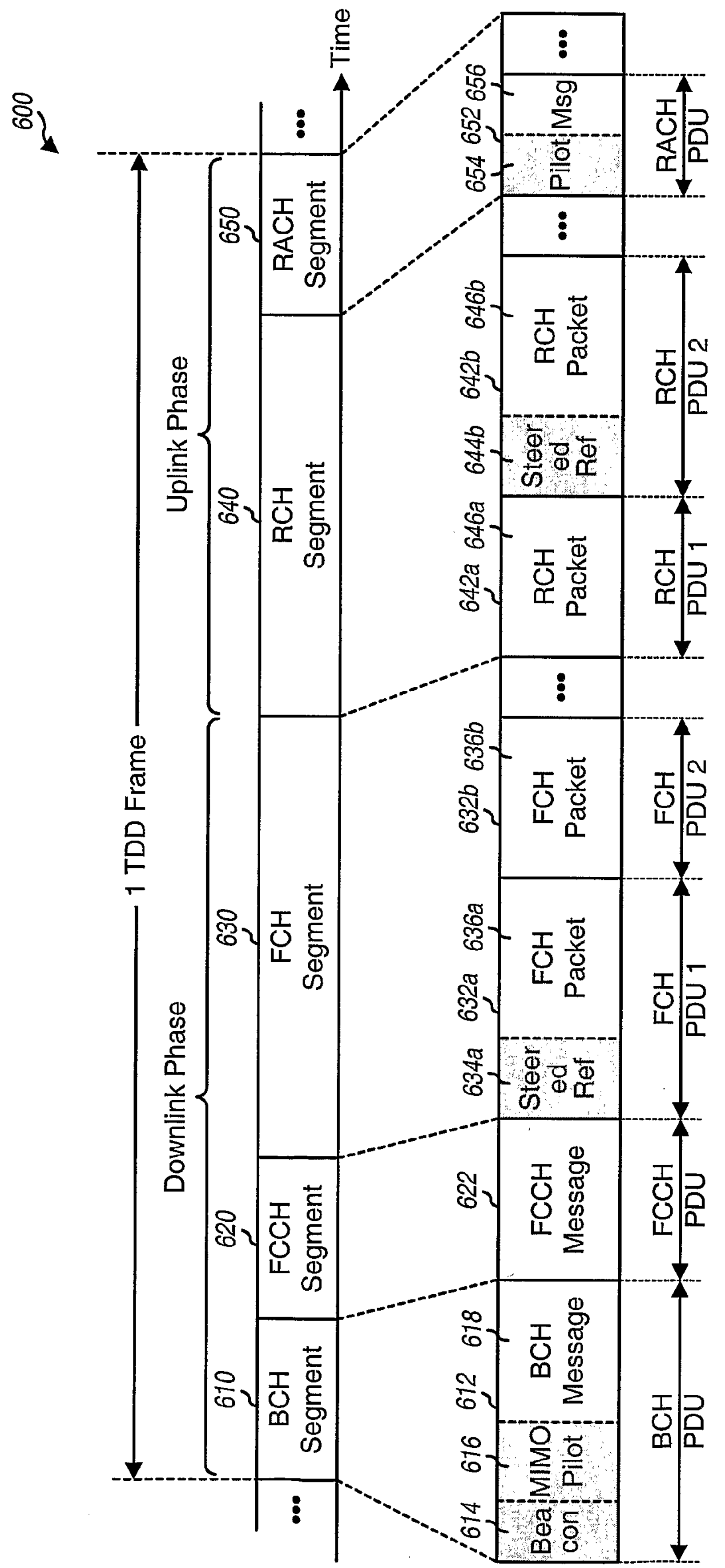
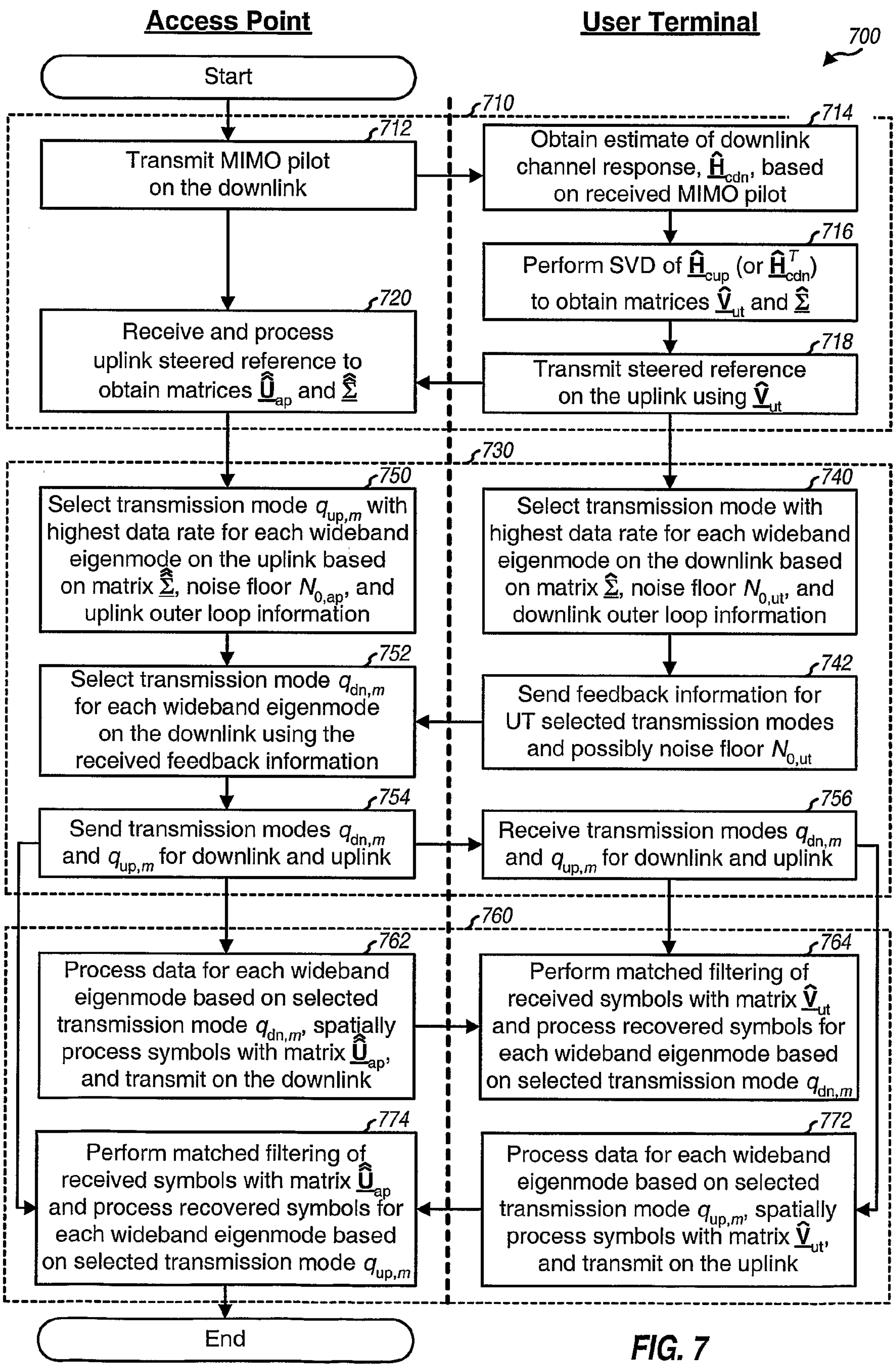

600
1 TDD Frame
Downlink Phase
Uplink Phase
610
BCH Segment
620
FCCH Segment
630
FCH Segment
640
RCH Segment
650
RACH Segment
Time
614
616
612
618
Bea con
MIMO Pilot
BCH Message
622
FCCH Message
634a
632a
636a
Steer ed Ref
FCH Packet
632b
636b
FCH Packet
642a
646a
RCH Packet
644b
642b
646b
Steer ed Ref
RCH Packet
654
652
656
Pilot
Msg
BCH PDU
FCCH PDU
FCH PDU 1
FCH PDU 2
RCH PDU 1
RCH PDU 2
RACH PDU

FIG. 6



Access Point
User Terminal
700
Start
710
712
Transmit MIMO pilot on the downlink
714
Obtain estimate of downlink channel response, $\hat{\underline{H}}_{cdn}$, based on received MIMO pilot
716
Perform SVD of $\hat{\underline{H}}_{cup}$ (or $\hat{\underline{H}}^{T}_{cdn}$) to obtain matrices $\hat{\underline{V}}_{ut}$ and $\hat{\underline{\Sigma}}$
720
Receive and process uplink steered reference to obtain matrices $\hat{\underline{U}}_{ap}$ and $\hat{\underline{\Sigma}}$
718
Transmit steered reference on the uplink using $\hat{\underline{V}}_{ut}$
730
750
Select transmission mode $q_{up,m}$ with highest data rate for each wideband eigenmode on the uplink based on matrix $\hat{\underline{\Sigma}}$, noise floor $N_{0,ap}$, and uplink outer loop information
740
Select transmission mode with highest data rate for each wideband eigenmode on the downlink based on matrix $\hat{\underline{\Sigma}}$, noise floor $N_{0,ut}$, and downlink outer loop information
752
Select transmission mode $q_{dn,m}$ for each wideband eigenmode on the downlink using the received feedback information
742
Send feedback information for UT selected transmission modes and possibly noise floor $N_{0,ut}$
754
Send transmission modes $q_{dn,m}$ and $q_{up,m}$ for downlink and uplink
756
Receive transmission modes $q_{dn,m}$ and $q_{up,m}$ for downlink and uplink
760
762
Process data for each wideband eigenmode based on selected transmission mode $q_{dn,m}$, spatially process symbols with matrix $\hat{\underline{U}}_{ap}$, and transmit on the downlink
764
Perform matched filtering of received symbols with matrix $\hat{\underline{V}}_{ut}$ and process recovered symbols for each wideband eigenmode based on selected transmission mode $q_{dn,m}$
774
Perform matched filtering of received symbols with matrix $\hat{\underline{U}}_{ap}$ and process recovered symbols for each wideband eigenmode based on selected transmission mode $q_{up,m}$
772
Process data for each wideband eigenmode based on selected transmission mode $q_{up,m}$, spatially process symbols with matrix $\hat{\underline{V}}_{ut}$, and transmit on the uplink
End

FIG. 7

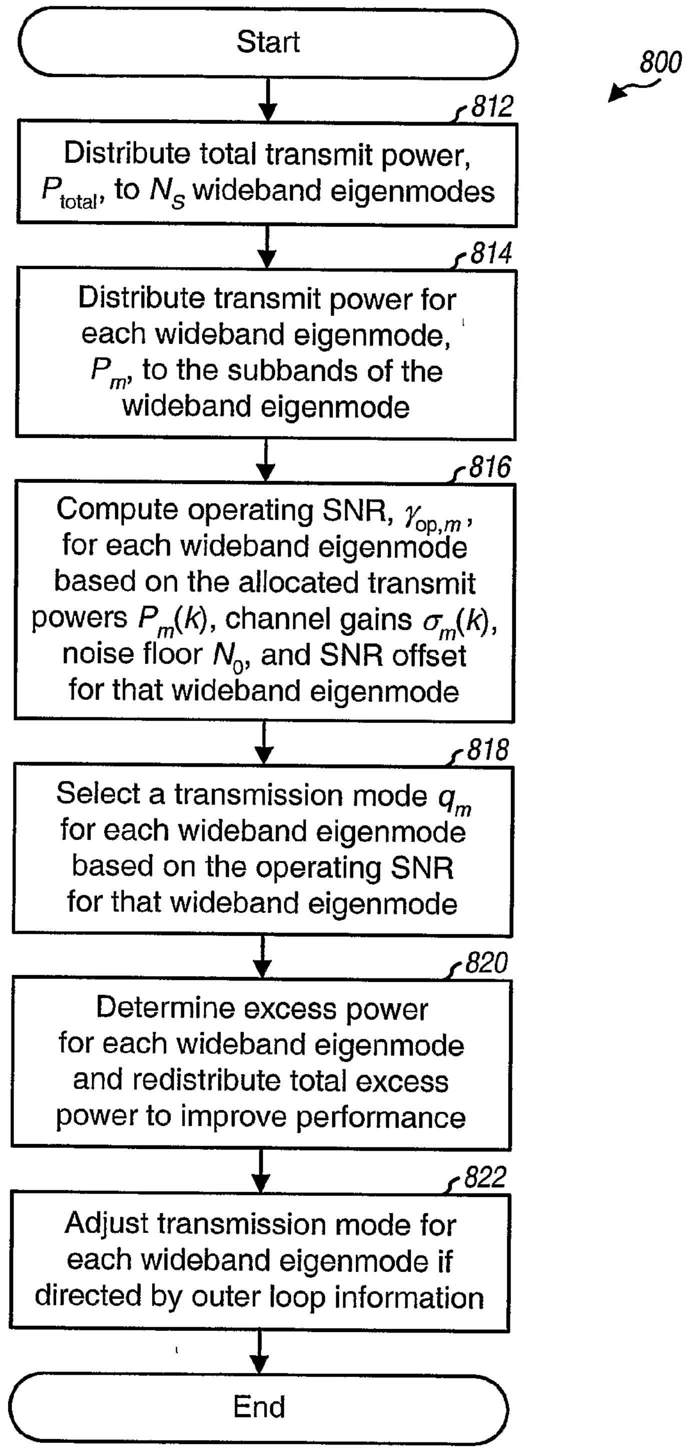

Start
800
812
Distribute total transmit power, P_{total}, to N_S wideband eigenmodes
814
Distribute transmit power for each wideband eigenmode, P_m, to the subbands of the wideband eigenmode
816
Compute operating SNR, $\gamma_{op,m}$, for each wideband eigenmode based on the allocated transmit powers $P_m(k)$, channel gains $\sigma_m(k)$, noise floor N_0, and SNR offset for that wideband eigenmode
818
Select a transmission mode q_m for each wideband eigenmode based on the operating SNR for that wideband eigenmode
820
Determine excess power for each wideband eigenmode and redistribute total excess power to improve performance
822
Adjust transmission mode for each wideband eigenmode if directed by outer loop information
End

FIG. 8

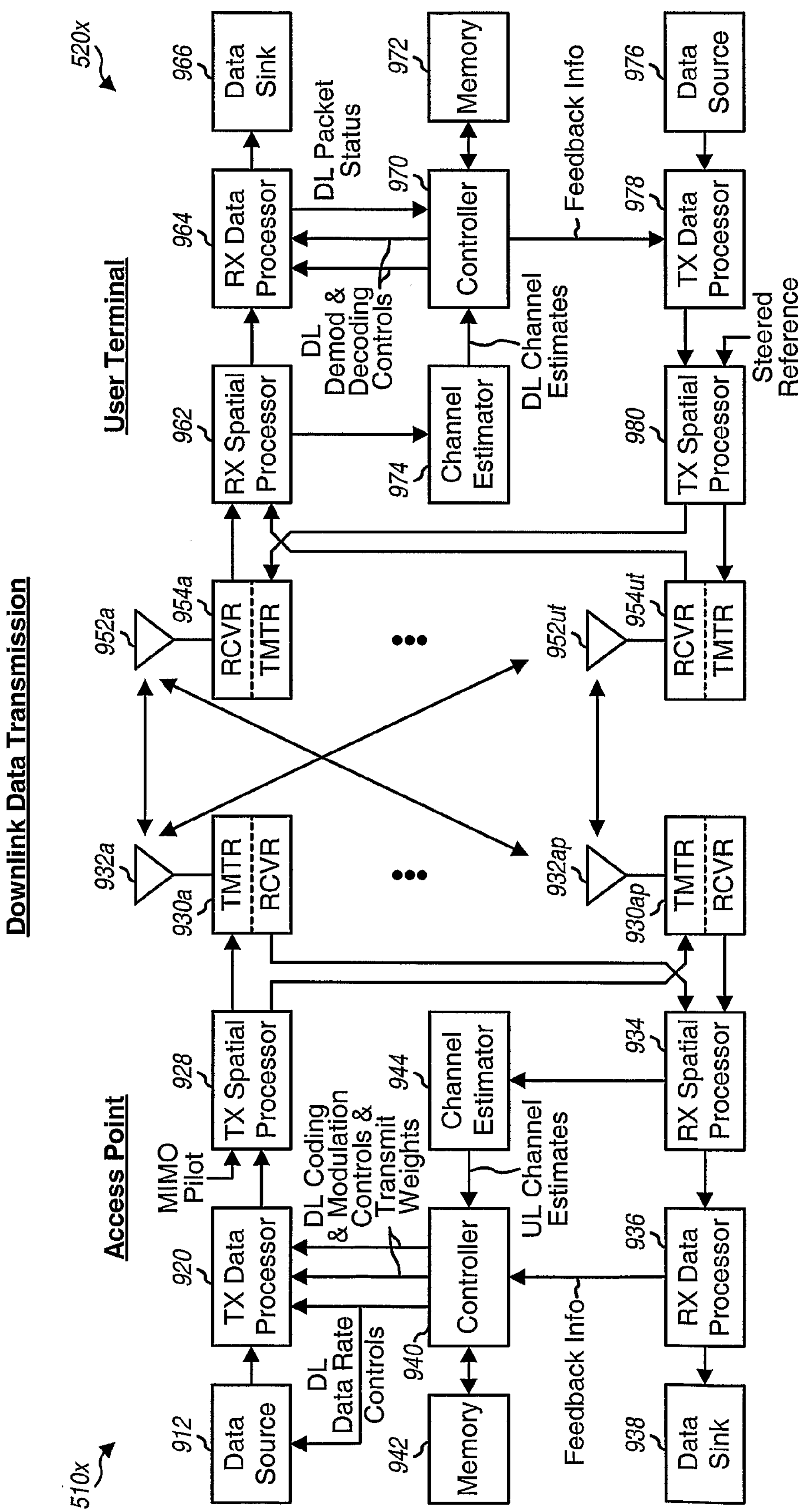

Downlink Data Transmission
Access Point
User Terminal
510x
520x
912
Data Source
920
TX Data Processor
928
TX Spatial Processor
MIMO Pilot
930a
TMTR
RCVR
932a
930ap
932ap
942
Memory
940
Controller
944
Channel Estimator
DL Data Rate Controls
DL Coding & Modulation Controls & Transmit Weights
UL Channel Estimates
Feedback Info
934
RX Spatial Processor
936
RX Data Processor
938
Data Sink
952a
954a
RCVR
TMTR
952ut
954ut
962
RX Spatial Processor
964
RX Data Processor
966
Data Sink
974
Channel Estimator
970
Controller
972
Memory
DL Packet Status
DL Demod & Decoding Controls
DL Channel Estimates
Feedback Info
980
TX Spatial Processor
978
TX Data Processor
976
Data Source
Steered Reference

FIG. 9A

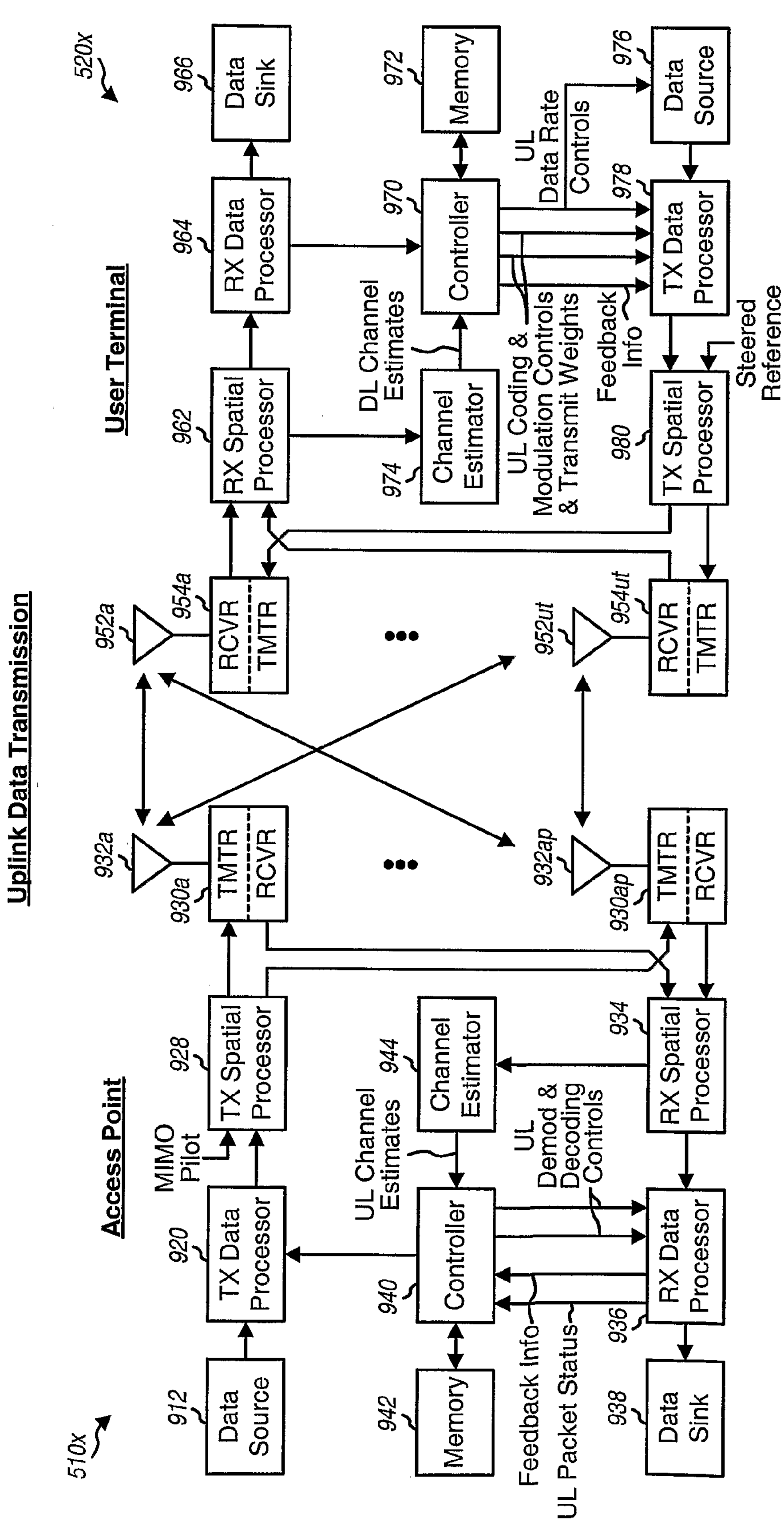

Uplink Data Transmission
Access Point
User Terminal
510x
520x
912
Data Source
920
TX Data Processor
MIMO Pilot
928
TX Spatial Processor
930a
TMTR
RCVR
932a
930ap
932ap
942
Memory
940
Controller
944
Channel Estimator
UL Channel Estimates
UL Demod & Decoding Controls
Feedback Info
UL Packet Status
938
Data Sink
936
RX Data Processor
934
RX Spatial Processor
952a
954a
952ut
954ut
962
RX Spatial Processor
964
RX Data Processor
966
Data Sink
974
Channel Estimator
DL Channel Estimates
970
Controller
972
Memory
UL Coding & Modulation Controls & Transmit Weights
UL Data Rate Controls
Feedback Info
978
TX Data Processor
976
Data Source
980
TX Spatial Processor
Steered Reference

FIG. 9B

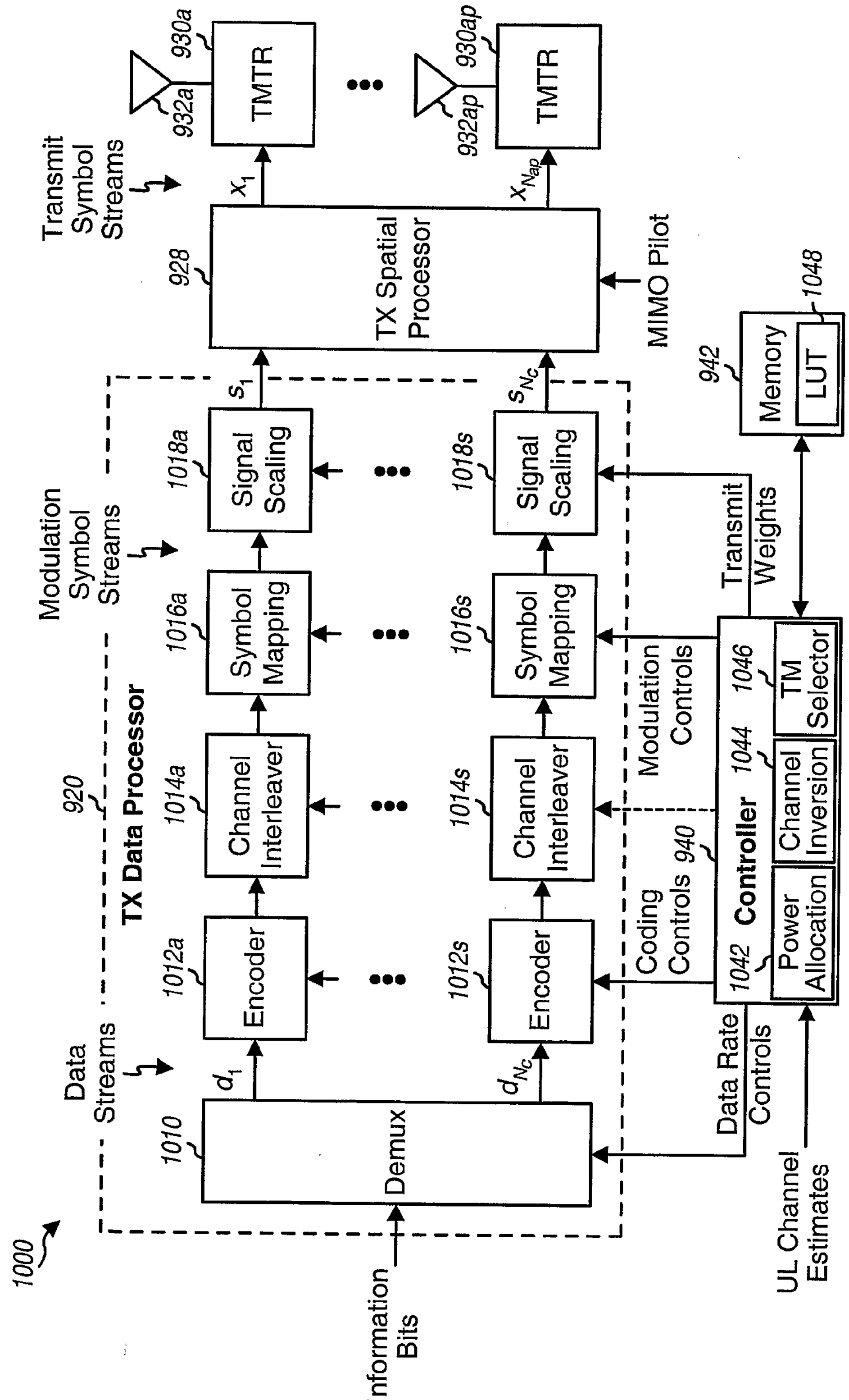

1000
Information Bits
TX Data Processor
920
1010
Demux
Data Streams
d_1
d_{N_c}
1012a
Encoder
1012s
Encoder
1014a
Channel Interleaver
1014s
Channel Interleaver
1016a
Symbol Mapping
1016s
Symbol Mapping
Modulation Symbol Streams
1018a
Signal Scaling
1018s
Signal Scaling
s_1
s_{N_c}
928
TX Spatial Processor
MIMO Pilot
Transmit Symbol Streams
x_1
$x_{N_{ap}}$
930a
TMTR
932a
930ap
TMTR
932ap
Coding Controls
Modulation Controls
Transmit Weights
Data Rate Controls
940
Controller
1042
Power Allocation
1044
Channel Inversion
1046
TM Selector
942
Memory
1048
LUT
UL Channel Estimates

FIG. 10

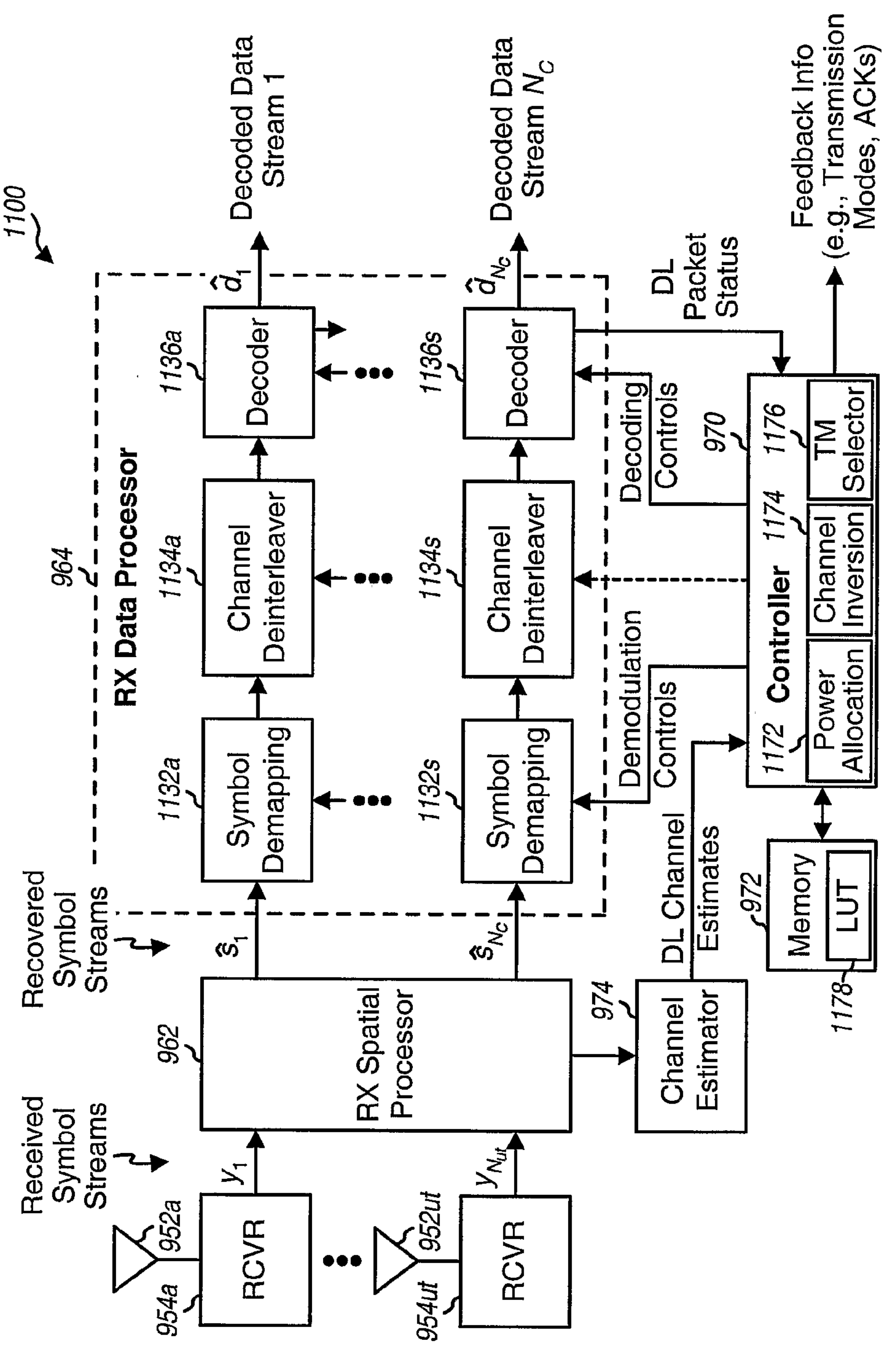

1100
Received Symbol Streams
Recovered Symbol Streams
952a
954a
RCVR
952ut
954ut
RCVR
y_1
$y_{N_{ut}}$
962
RX Spatial Processor
$\hat{s}_1$
$\hat{s}_{N_C}$
974
Channel Estimator
DL Channel Estimates
964
RX Data Processor
1132a
Symbol Demapping
1132s
Symbol Demapping
1134a
Channel Deinterleaver
1134s
Channel Deinterleaver
1136a
Decoder
1136s
Decoder
$\hat{d}_1$
$\hat{d}_{N_C}$
Decoded Data Stream 1
Decoded Data Stream N_C
Demodulation Controls
Decoding Controls
DL Packet Status
970
Controller
1172
Power Allocation
1174
Channel Inversion
1176
TM Selector
Feedback Info (e.g., Transmission Modes, ACKs)
972
Memory
LUT
1178

FIG. 11

Downlink Rate Control

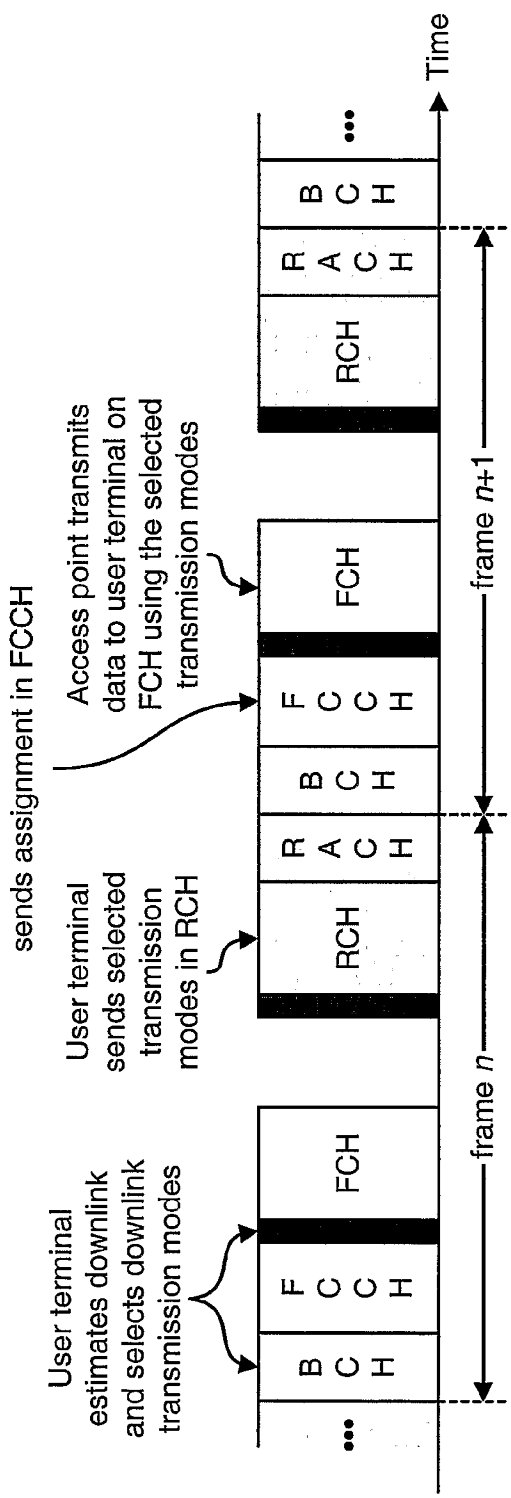


FIG. 12A

Uplink Rate Control

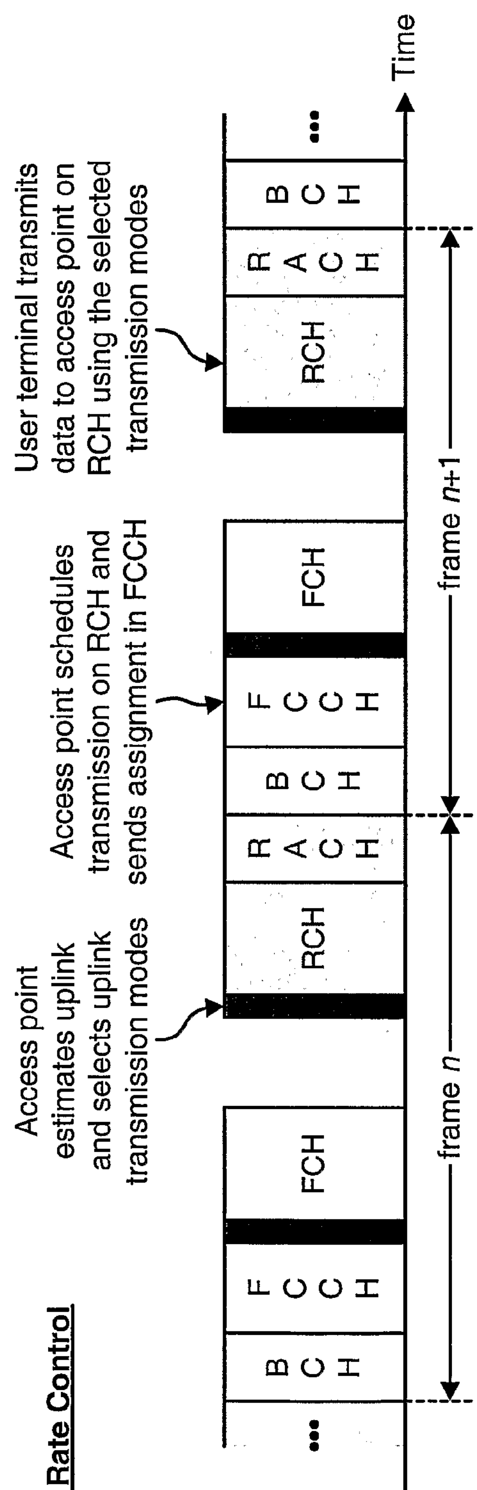


FIG. 12B

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(81) **Designated States** *(unless otherwise indicated, for every kind of national protection available)*: AE, AG, AL, AM, AT, AU, AZ, BA, BB, BG, BR, BW, BY, BZ, CA, CH, CN, CO, CR, CU, CZ, DE, DK, DM, DZ, EC, EE, EG, ES, FI, GB, GD, GE, GH, GM, HR, HU, ID, IL, IN, IS, JP, KE, KG, KP, KR, KZ, LC, LK, LR, LS, LT, LU, LV, MA, MD, MG, MK, MN, MW, MX, MZ, NA, NI, NO, NZ, OM, PG, PH, PL, PT, RO, RU, SC, SD, SE, SG, SK, SL, SY, TJ, TM, TN, TR, TT, TZ, UA, UG, US, UZ, VC, VN, YU, ZA, ZM, ZW.

(84) **Designated States** *(unless otherwise indicated, for every kind of regional protection available)*: ARIPO (BW, GH, GM, KE, LS, MW, MZ, NA, SD, SL, SZ, TZ, UG, ZM, ZW), Eurasian (AM, AZ, BY, KG, KZ, MD, RU, TJ, TM), European (AT, BE, BG, CH, CY, CZ, DE, DK, EE, ES, FI, FR, GB, GR, HU, IE, IS, IT, LU, MC, NL, PL, PT, RO, SE, SI, SK, TR), OAPI (BF, BJ, CF, CG, CI, CM, GA, GN, GQ, GW, ML, MR, NE, SN, TD, TG).

Published:
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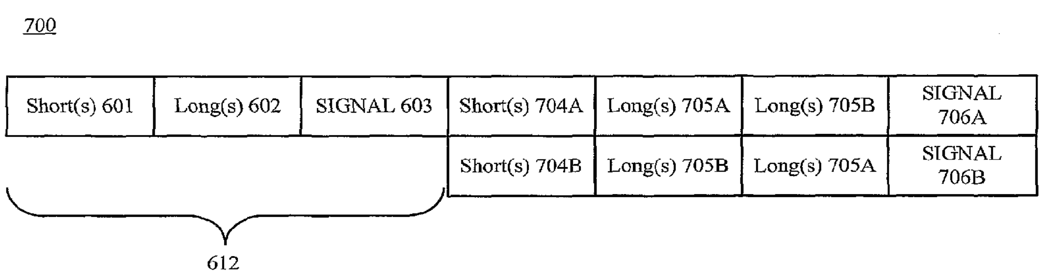
For two-letter codes and other abbreviations, refer to the "Guidance Notes on Codes and Abbreviations" appearing at the beginning of each regular issue of the PCT Gazette.

(54) Title: MULTIPLE-INPUT MULTIPLE-OUTPUT SYSTEM AND METHOD



(57) Abstract: A multiple-input multiple-output (MIMO) system can transmit on multiple antennas simultaneously and receive on multiple antennas simultaneously. Unfortunately, because a legacy 802.11a/g device is not able to decode multiple data streams, such a legacy device may "stomp" on a MIMO packet by transmitting before the transmission of the MIMO packet is complete. Therefore, MIMO systems and methods are provided herein to allow legacy devices to decode the length of a MIMO packet and to restrain from transmitting during that period. These MIMO systems and methods are optimized for efficient transmission of MIMO packets.

WO 2005/046113 A2

Multiple-Input Multiple-Output System And Method

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Won-Joon Choi
Jeffrey M. Gilbert
Ardavan Maleki Tehrani

RELATED APPLICATIONS

[0001] This application claims priority of U.S. Provisional Patent Application 60/517,445, entitled "Method For Multiple Input Multiple Output Systems" filed November 4, 2003.

BACKGROUND OF THE INVENTION

Field of the Invention

[0002] The present invention relates to a multiple-input multiple-output (MIMO) system and method in a wireless communication environment, and in one embodiment to a MIMO method and system that facilitates backwards compatibility with legacy devices.

Description of the Related Art

[0003] The design of communication systems for wireless local area networks (WLANs) is based on a family of standards described in IEEE 802.11. For example, the 802.11a specification provides up to 54 Mbps in the 5 GHz band whereas the 802.11g specification also provides up to 54 Mbps but in the 2.4 GHz band. Both the 802.11a/g specifications use an orthogonal frequency division multiplexing (OFDM) encoding scheme.

[0004] Notably, the 802.11a/g specifications provide for only one data stream being transmitted or received at any given time. For example, Figure 1 illustrates a simplified system 100 including a transmitter 101 that

can provide a single output at any given time and a receiver 102 that can process a single input at any given time. Thus, system 100 is characterized as a single input single output system.

[0005] To address multipath and, more particularly, the fading caused by multipath (wherein objects in the environment can reflect a transmitted wireless signal) and other conditions, a wireless system can employ various techniques. One such technique is switch diversity, wherein transmitters and/or receivers can selectively switch between multiple antennas. For example, Figure 2 illustrates a simplified system 200 in which transmitter 101 can choose to send signals from antenna 201A or antenna 201B (using a switch 203) whereas receiver 102 can choose to process signals from antenna 202A or antenna 202B (using a switch 204). Thus, system 200 is characterized as a switched diversity antenna configuration.

[0006] Figure 3 illustrates a simplified multiple-input multiple-output (MIMO) system 300, which can transmit on multiple antennas simultaneously and receive on multiple antennas simultaneously. Specifically, a transmitter 301 can transmit signals simultaneously from antenna 302A (using a transmitter chain 303A) and from antenna 302B (using a transmitter chain 303B). Similarly, a receiver 304 can receive signals simultaneously from antenna 305A (using a receiver chain 306A) and from antenna 305B (using a receiver chain 306B).

[0007] Note that there are a number of types of MIMO systems. For example, MIMO-AG refers to a MIMO system compatible with both 802.11a and 802.11g. In contrast, MIMO-SM refers to a MIMO system with spatial

multiplexing. The use of the acronym "MIMO" hereinafter refers to MIMO-SM.

[0008] The use of multiple antennas, depending on the specific implementation, can either extend the range or increase the data rate at a given range. For example, Figure 4 illustrates the median data rates for various antenna configurations over relative distances. Waveform 401 represents a single antenna configuration; waveform 402 represents a switched diversity antenna configuration; and waveform 403 represents a MIMO antenna configuration. Notably, at any relative distance between 2 and 4, the median data rate for the MIMO antenna configuration is significantly greater than the median data rates for either the single antenna configuration or the switched diversity antenna configuration. For example, at relative distance 3, which represents a top end for a typical home space 404, the median data rate for a MIMO antenna configuration (50 Mbps) is significantly greater than the median data rates for a single antenna configuration (18 Mbps) or even for a switched diversity antenna configuration (33 Mbps).

[0009] A MIMO system can also advantageously minimize the differences in signal to noise ratio (SNR) across different frequency bins. For example, Figure 5 illustrates the SNRs for various antennas across various frequency bins, i.e. SNRs 501 for a first antenna (waveform represented by the dotted line), SNRs 502 for a second antenna (waveform represented by the dashed line), and SNRs 503 for simultaneous usage of the first and second antennas (waveform represented by the solid line). Note that both SNRs 501 and 502 can vary significantly over frequency bins 0-60. In contrast, a MIMO system simultaneously using both the first and second antennas, shown by SNRs 503, can minimize the differences in SNR

across different frequency bins (i.e. notches on one channel are compensated for by non-notches in the other channel), thereby allowing more effective compensation for such SNR in the receiver chains and/or transmitter chains.

[0010] In MIMO system 300 (Figure 3), receiver 304 uses multiple chains (i.e. chains 306A and 306B) to receive and decode the multiple data streams (e.g. packets) transmitted by transmitter 301. Unfortunately, because a legacy 802.11a/g device is not able to decode multiple data streams, such a legacy device may "stomp" on a MIMO packet by transmitting before the transmission of the MIMO packet is complete.

[0011] Therefore, a need arises for a MIMO system and method that allows legacy devices to decode the length of a MIMO packet and to restrain from transmitting during that period. A further need arises for an efficient way to transmit MIMO packets.

SUMMARY OF THE INVENTION

[0012] A multiple-input multiple-output (MIMO) system can transmit on multiple antennas simultaneously and receive on multiple antennas simultaneously. Unfortunately, because a legacy 802.11a/g device is not able to decode multiple data streams, such a legacy device may "stomp" on a MIMO packet by transmitting before the transmission of the MIMO packet is complete. Therefore, MIMO systems and methods are provided herein to allow legacy devices to decode the length of a MIMO packet and to restrain from transmitting during that period. These MIMO systems and methods are optimized for efficient transmission of MIMO packets.

[0013] For example, a time-division training pattern for MIMO packets is provided. In this pattern, a first

antenna can transmit a short symbol, a first long symbol, and then a legacy SIGNAL symbol. A second antenna can transmit a second long symbol after transmission of the legacy SIGNAL symbol. The first and second antennas can transmit SIGNAL symbols (associated with MIMO data) substantially simultaneously after transmission of the second long symbol.

[0014] Another pattern for MIMO packets is provided. In this pattern, a short symbol can be transmitted by a first antenna and a second antenna. The short symbol can be split between a predetermined set of short bins. Notably, the first antenna can be associated with a first set of short bins whereas the second antenna can be associated with a second set of short bins. A long symbol can be transmitted substantially simultaneously by the first and second antennas after transmission of the second short symbol. The long symbol can be associated with a first set of long bins and a second set of long bins. Notably, the first antenna can transmit using the first set of long bins before using the second set of long bins. In contrast, the second antenna can transmit using the second set of long bins before using the first set of long bins. SIGNAL symbols associated with multiple-input multiple-output data can be transmitted substantially simultaneously by the first and second antennas.

[0015] In one embodiment, the first set of short bins can include -24, -16, -8, 4, 12, 20 and the second set of short bins can include -20, -12, -4, 8, 16, 24. In another embodiment, the first set of short bins can include -24, -16, -8, 8, 16, 24 and the second set of short bins can include -20, -12, -4, 4, 12, 20.

[0016] In one embodiment, the first set of long bins can include -26, -24, … -2, 1, 3, … 25 and the second set

of long bins can include -25, -23, … -1, 2, 4, … 26. In another embodiment, the first set of long bins can include -26, -24, … -2, 2, 4, … 26, and the second set of long bins can include -25, -23, … -1, 1, 3, … 25.

[0017] In one embodiment, the pattern can further include computing peak-to-average ratio (PAR) values for at least two split patterns of the short bins and using the split pattern having a lowest PAR value. In another embodiment, the pattern can further include computing peak-to-average ratio (PAR) values for at least two split patterns of the long bins and using the split pattern having an optimized PAR value.

[0018] The first and second sets of short bins can use a different frequency shift. For example, if the pattern is using 1 out of every N bins, then a frequency shift pattern can include 1 up to N-1 bins.

[0019] In one embodiment, the first antenna can be implemented using a set of antennas. In this case, complex weights can be applied across bins of the first set of antennas, thereby mitigating beam-forming effects. The complex weights include at least one of phase shifts or phase magnitudes, and wherein mitigating beam-forming effects creates a substantially omni-directional transmission.

[0020] In one embodiment having a legacy header, the pattern can further include an encoding symbol transmitted after the legacy header to indicate that a MIMO packet is being transmitted. The encoding symbol can indicate at least a number of transmitted data streams. In one embodiment, the encoding symbol can include the SIGNAL symbols associated with MIMO data. These SIGNAL symbols can include flipped pilot tones, wherein the flipped pilot tones are different than that

of regular symbols that would otherwise appear in that location.

[0021] A method of sending a MIMO packet in a legacy device environment is provided. In this method, a reserved set of bits in a legacy SIGNAL symbol can be set to a predetermined value, thereby indicating a multiple-input multiple-output signal is being transmitted. In another method, a set of bits in the legacy SIGNAL symbol can indicate information associated with a MIMO packet. In one embodiment, the set of bits can include a plurality of least significant bits of a length field of the legacy SIGNAL symbol. The information associated with the MIMO packet could indicate the number of transmitted data streams associated with the MIMO packet. In another method, a 'modulo' operation can be performed on a set of bits in the legacy SIGNAL symbol to indicate information (e.g. the number of streams) associated with the MIMO packet.

[0022] A method of tracking and correcting phase variations of multiple received data symbols for a MIMO signal is provided. In this method, a plurality of pilot bins can be inserted into each data symbol. In one embodiment, phase shifting can be added using a pattern across the plurality of pilot bins. For example, the pattern of the phase shifting can be rotated (e.g. cyclically) across the plurality of pilot bins. In one embodiment, four pilot bins can be inserted into each data symbol in a format of $[1\ 1\ 1\ -1]*p_l$, wherein $[1\ 1\ 1\ -1]$ is a pattern across the four pilot bins and p_l is a pilot polarity for symbol *l*.

[0023] Another method of tracking and correcting phase variations of multiple received data symbols for a MIMO signal is also provided. In this method, orthogonal patterns can be provided across data streams over any

interval of M data symbols long. Providing orthogonal patterns can conform to the equation:

$$\frac{1}{M}\sum_{l=k}^{k+M-1} q_m(l)q_n^*(l) = \delta_{mn},$$

wherein M represents a number of transmitted data streams, m represents a stream, k represents a starting index of M orthogonal data symbols, l represented an index of MIMO symbols, and δ_{mn} is equal to 1 for m=n or equal to 0 for m≠n. For M transmitted data streams, then a modulating pattern for stream m, wherein $1 \le m \le M$ and $l \ge 0$, is $q_m(l) = e^{j\frac{2\pi}{M}(m-1)l}$.

[0024] A method of joint pilot tracking across streams is provided. In this method, a received signal in each pilot bin can be estimated based on a channel estimation and known pilot patterns. A received signal on a receiver n in pilot K is represented by

$$y_{n,k} = \sum_m H_{n,m,k} e^{j\theta} \cdot s_{m,k} + n_{n,k}$$

wherein $s_{m,k}$ is a pilot symbol of stream m, θ is a common phase offset, $H_{n,m,k}$ is a channel response, and $n_{n,k}$ is noise, wherein a common phase offset is represented by

$$\theta = \text{angle}\left(\sum_{n,k} y_{n,k} \cdot \left(\sum_m \hat{H}_{n,m,k} s_{m,k}\right)^*\right)$$

wherein $\hat{H}_{n,m,k}$ is the channel estimation.

[0025] A method of pilot tracking per transmit chain is provided. In this method, the MIMO detection algorithms can be applied to pilot bins to detect the pilots $\hat{s}_{m,k}$, wherein $\hat{s}_{m,k} \approx s_{m,k} \cdot e^{j\theta_t(m)}$, where $\theta_t(m)$ is a phase offset of stream m. A phase difference can be averaged between decoded pilots and ideal pilots over the pilot

bins of each data stream to generate a phase estimate $\hat{\theta}_t(m) = \text{angle}(\sum_k \hat{s}_{m,k} \cdot s^*_{m,k})$.

[0026] A method of pilot tracking per transmit/receive chains is provided. This method can include modulating pilot polarity sequences with orthogonal patterns, thereby estimating phase separately for each transmit/receive chain. If a number of transmitted data streams is M, then a modulating pattern for stream m, wherein $1 \le m \le M$, can be represented by $q_m(l) = e^{j\frac{2\pi}{M}(m-1)l}$, where $l \ge 0$ is the index of the MIMO symbols. The method can further include estimating a phase offset of stream m on a receive antenna n by averaging over a plurality of pilot bins, represented by $\theta_{n,m} = \text{angle}(\sum_k v_{n,m,k}) = \text{angle}(\sum_k \sum_l y_{n,k}(l) \cdot r^*_{m,k}(l) \cdot H^*_{n,m,k})$.

[0027] A method of splitting source data bits to form a MIMO signal is provided. In this method, bits can be added to the source data bits to initialize and terminate an encoder, thereby creating modified source data bits. The modified source data bits can be provided to the encoder, thereby creating encoded source data bits. The encoded source data bits can then be split into N data streams.

[0028] Another method of splitting source data bits to form a MIMO signal is provided. In this method, the source data bits can be split into N data streams. Bits can be added to the N data streams to initialize and terminate N encoders, thereby creating N modified data streams. The method can further include selecting a total number of bits such that when split across symbols for each of the N data streams, the number of symbols in each data stream is substantially equal.

[0029] Yet another method of splitting source data bits to form a MIMO signal is provided. In this method, bits can be added to source data bits to initialize and terminate an encoder, thereby creating modified source data bits. The modified source data bits can be provided to the encoder, thereby creating encoded source data bits. The encoded source data bits can be provided to a puncturer, thereby creating punctured source data bits. Then, the punctured source data bits can be split into N data streams.

[0030] A method of indicating a length of a MIMO packet using a legacy SIGNAL symbol is provided. This legacy SIGNAL symbol can include a rate field and a length field. However, the length of the MIMO packet may be longer than can be represented using the length field. In this case, the method can include using the rate field as well as the length field to represent the length of the MIMO packet. For example, a pseudo-rate value can be provided in the rate field and a pseudo-length value can be provided in the length field. In one embodiment, the pseudo-rate value can be the lowest legacy rate and the pseudo-length value can be an actual legacy length representing a transmit duration. In another embodiment, a MIMO SIGNAL symbol of the MIMO packet includes a relative packet length.

[0031] A pattern for MIMO packets is provided. The pattern can include a legacy header and a MIMO header. The legacy header can include a plurality of short symbols for determining automatic gain control for receipt of the legacy header. In contrast, the MIMO header can include a second plurality of short symbols for facilitating automatic gain control for receipt of the MIMO header.

[0032] Another pattern for MIMO packets is provided. This pattern can include a first short symbol transmitted by a plurality of antennas. Notably, the first short symbol can be split between a predetermined set of short bins, wherein each of the plurality of antennas can be associated with a subset of the short bins. The first short symbol can be used for automatic gain control for a MIMO packet (the MIMO packet including the first short symbol).

[0033] The pattern can further include a first long symbol transmitted substantially simultaneously by the plurality of antennas. Notably, the first long symbol can be associated with sets of long bins, wherein each antenna transmits using a different order of the sets of long bins. The first long symbol can be used for MIMO channel estimation (the MIMO packet further includes a first long symbol).

[0034] In one embodiment, the plurality of antennas includes a first and second antennas. In this case, the first and second antennas can transmit the first short symbol after transmission of the legacy SIGNAL symbol. The first antenna can be associated with a first set of short bins whereas the second antenna can be associated with a second set of short bins. The first and second antennas can transmit the first long symbol substantially simultaneously. Notably, the first long symbol can be associated with a first set of long bins and a second set of long bins, wherein the first antenna transmits using the first set of long bins before using the second set of long bins, and the second antenna transmits using the second set of long bins before using the first set of long bins.

[0035] The pattern can also include SIGNAL symbols associated with MIMO transmitted substantially

simultaneously by the first and second antennas after the first short symbol and the first long symbol. The pattern can further include a second short symbol, a second long symbol, and a legacy SIGNAL symbol. The second short symbol can be used for automatic gain control for a legacy header. The legacy header can include the second short symbol, the second long symbol, and the legacy SIGNAL symbol. Notably, the legacy header is transmitted before the MIMO header.

[0036] Yet another pattern for MIMO packets is provided. This pattern can also include a legacy header and a MIMO header. The legacy header can include a first plurality of long symbols used for legacy device channel estimation. The MIMO header can include a second plurality of long symbols used for MIMO device channel estimation.

[0037] Yet another pattern for multiple-input multiple-output (MIMO) packets is provided. This pattern can include a first long symbol transmitted by a plurality of antennas. The first long symbol can be transmitted substantially simultaneously by the plurality of antennas. Notably, the first long symbol can be associated with sets of long bins, wherein each antenna transmits using a different order of the sets of long bins. The first long symbol can be used for MIMO channel estimation for a MIMO packet (the MIMO packet including the first long symbol).

[0038] This pattern can further include a first short symbol. The first short symbol can also be transmitted by the plurality of antennas. Notably, the first short symbol can be split between a predetermined set of short bins, wherein each of the plurality of antennas is associated with a subset of the short bins. The first short symbol can be used for automatic gain control for

the MIMO packet (the MIMO packet including the first short symbol).

[0039] In one embodiment, the plurality of antennas can include first and second antennas. The first and second antennas can transmit the first short symbol after transmission of the legacy SIGNAL symbol. The first antenna can be associated with a first set of short bins and the second antenna can be associated with a second set of short bins. The first and second antennas can transmit the first long symbol substantially simultaneously. The first long symbol can be associated with a first set of long bins and a second set of long bins. Notably, the first antenna can transmit using the first set of long bins before using the second set of long bins. In contrast, the second antenna can transmit using the second set of long bins before using the first set of long bins.

[0040] The pattern can further include SIGNAL symbols associated with MIMO transmitted substantially simultaneously by the first and second antennas after the first short symbol and the first long symbol.

[0041] A method of decoding a plurality of encoded data streams for a MIMO transmission is provided. In this method, for decoding, the data bits from good bins can be weighted more heavily than the data bits from bad bins. For example, the bin weights can be proportional to a signal to noise ratio (SNR) or to a square root of the SNR.

[0042] The weighting can influence Viterbi branch metrics computation. In one embodiment, the method can further include determining the impact of error propagation based on the following equations for computing effective noise terms for second and third streams:

$$\tilde{\sigma}_2^2 = \sigma_2^2 + \left|w_2^* h_1\right|^2 \cdot \sigma_1^2$$
$$\tilde{\sigma}_3^2 = \sigma_3^2 + \left|w_3^* h_2\right|^2 \cdot \tilde{\sigma}_2^2 + \left|w_3^* h_1\right|^2 \cdot \sigma_1^2$$

wherein σ_m^2 is an original noise term, w_m is a nulling vector, h_m is a channel, and $\tilde{\sigma}_m^2$ is an effective noise term for an *m*-th data stream.

[0043] A method for modifying channel corrections for a plurality of receiver chains is provided. In this method, channel estimates for the plurality of receiver chains can be received. Gain adjustment values for the plurality of receiver chains can be computed based on a noise floor and automatic gain control values. Then, the gain adjustment values can be applied to the plurality of receiver chains.

[0044] A method of using phase estimates for a MIMO system is provided. In this method, a single joint phase estimate can be used from a plurality of data streams to compute a phase correction applicable to all data streams. In one embodiment, the plurality of data streams includes all data streams.

[0045] A method of providing phase estimations for each transmit/receive pair is provided. In this method, a phase offset of each element of a channel matrix *H*, $\theta_{n,m}(1 \le m \le M, 1 \le n \le N)$, can be estimated from pilots and converted the phase offset into $\theta_t(m)(1 \le m \le M)$ and $\theta_r(n)(1 \le n \le N)$. In channel matrix H,

$$\begin{bmatrix} 1_N & & & & I_N \\ & 1_N & & & I_N \\ & & \ddots & & \vdots \\ & & & 1_N & I_N \end{bmatrix} \begin{bmatrix} \theta_t(1) \\ \theta_t(2) \\ \vdots \\ \theta_t(M) \\ \theta_r \end{bmatrix} = \begin{bmatrix} \theta_1 \\ \theta_2 \\ \vdots \\ \theta_M \end{bmatrix} \Leftrightarrow A \cdot \Theta_1 = \Theta_2 \qquad \Rightarrow \qquad \Theta_1 = pinv(A) \cdot$$

wherein 1_N is an N-by-1 vector of all 1's, I_N is an identity matrix of size N, $\theta_r = [\theta_r(1) \quad \theta_r(2) \quad \ldots \quad \theta_r(N)]^T$ is a phase vector at N receivers, and $\theta_m = [\theta_{1,m} \quad \theta_{2,m} \quad \ldots \quad \theta_{N,m}]^T$ is a phase vector of an m-th column of matrix H.

[0046] A method of optimizing transmission of a MIMO signal is provided. In this method, a quality of a channel can be assessed using a packet received by an intended receiver from a transmitter of the MIMO signal. At this point, a packet (e.g. a CTS packet or an ACK packet) can be sent from the intended receiver to the transmitter, the packet including feedback information for optimizing transmission. This feedback information can be derived from a plurality of data streams previously transmitted substantially simultaneously. For example, the feedback information can include (1) channel estimates or (2) a detection pilot EVM computed from channel corrected pilots and known clean pilots.

[0047] In one embodiment, the feedback information can include a data rate to be used by the transmitter. In another embodiment, the feedback information can include an indicator for a minimum data rate, a maximum data rate, a higher data rate, and/or a lower data rate to be used by the transmitter.

[0048] A method of optimizing transmission of a transmitted MIMO signal is provided. In this method, a quality of a channel can be assessed using a MIMO packet, the MIMO packet being received by a transmitter for the MIMO signal from an intended receiver. Optimized transmit information can be determined based on the MIMO packet.

[0049] A method of determining receiver selection for a MIMO signal in a diversity antenna system is provided. At least one receiver chain is connectable to a plurality

of receive antennas. In this method, for each receiver chain, a receive antenna having a strongest signal can be selected.

[0050] A method of determining receiver selection for a MIMO signal in a diversity antenna system is also provided. In this method, possible combinations of receive antennas can be determined, wherein at least one receiver chain is connectable to a plurality of receive antennas. The signal to noise (SNR) can be computed for each combination. Then, the combination having a minimum SNR can be selected.

[0051] A method of selecting a split sequence is also provided. In this method, the power-to-average ratios (PARs) for a plurality of split sequences can be computed. Then, the split sequence having an optimized PAR can be selected.

[0052] The advantages of these MIMO systems and methods will now be described in reference to the following figures.

BRIEF DESCRIPTION OF THE FIGURES

[0053] Figure 1 illustrates a simplified system including a single input single output antenna configuration.

[0054] Figure 2 illustrates a simplified system including a switched diversity antenna configuration.

[0055] Figure 3 illustrates a simplified multiple-input multiple-output (MIMO) system, which can transmit on multiple antennas simultaneously and receive on multiple antennas simultaneously.

[0056] Figure 4 illustrates the median data rates for various antenna configurations over relative distances.

[0057] Figure 5 illustrates the SNRs for various antennas across various frequency bins.

[0058] Figure 6 illustrates an exemplary time-division training pattern for a MIMO packet including a legacy header.

[0059] Figure 7A illustrates an exemplary pattern for a MIMO packet including split short and long symbols to facilitate improved receiver gain control.

[0060] Figure 7B illustrates an exemplary split short and long symbols for three data streams.

[0061] Figure 7C illustrates the set of bits in the legacy SIGNAL symbol that can be set to indicate a MIMO packet as well as the number of transmitted data streams.

[0062] Figure 8 illustrates an exemplary shared encoder system for two spatial streams.

[0063] Figure 9 illustrates another exemplary shared encoder system for two spatial streams.

[0064] Figure 10 illustrates an exemplary separate encoder system for two spatial streams.

[0065] Figure 11 illustrates a portion of an exemplary receiver that can modify channel correction for a plurality of receiver chains.

[0066] Figure 12 illustrates the data rate of several transmitter/receiver antenna configurations over relative distances.

[0067] Figure 13 illustrates the data rate of various turbo and non-turbo antenna configurations over relative distances.

DETAILED DESCRIPTION OF THE FIGURES

Legacy Header And Symbol Splitting

[0068] In accordance with one embodiment, a legacy device can be prevented from "stomping" on MIMO signals (i.e. transmitting before the transmission of the MIMO packet is complete) by receiving a backward compatible preamble preceding the MIMO packet. This backward

compatible preamble, which is compatible with IEEE 802.11a/g systems, can advantageously allow the legacy device to decode the length of the MIMO packet and restrain from transmitting during that period. Additionally, this preamble can indicate whether the attached packet is a MIMO packet and, if it is, the number of data streams being transmitted.

[0069] Figure 6 illustrates an exemplary time-division training pattern 600 for a MIMO packet including this preamble. Specifically, a preamble 612, also called a legacy header herein, can include the standard 802.11a/g short symbol(s), long symbol(s) and SIGNAL symbol(s). Note that subsequent references to these symbols, although in plural form, could refer to either plural or single symbols.

[0070] In one embodiment, spatial streams 610 and 611 can be transmitted respectively from two (e.g. first and second) antennas. In other embodiments, spatial stream 610 can be transmitted from a plurality of antennas in a beam-forming antenna configuration. Thus, spatial stream 610 can be characterized as being transmitted by a set of antennas. For convenience, spatial stream 610 is described as being transmitted by a first antenna whereas spatial stream 611 is described as being transmitted by a second antenna.

[0071] In legacy header 612, short symbols 601 can be used for coarse ppm estimation and timing. Long symbols 602 transmitted from the first antenna can be used to estimate the channel from the first antenna. Notably, in one embodiment, SIGNAL symbol 603 can advantageously include the length information for the MIMO packet, thereby preventing a legacy device from stomping on the MIMO packet. Long symbols 604, which are transmitted from the second antenna but otherwise identical to long

symbols 602, can be used to estimate the channel from the second antenna. SIGNAL symbols 605A and 605B can include information regarding the modulation and length of the MIMO portions of spatial streams 610 and 611, respectively (wherein the MIMO portions are those portions after legacy header 612).

[0072] Figure 7A illustrates another exemplary pattern 700 for a MIMO packet including legacy header 612. Pattern 700 can advantageously split short and long symbols to facilitate improved receiver gain control (i.e. to ensure continuous received power even if the transmit paths are dissimilar). In pattern 700, additional short symbols 704A and 704B can be inserted after legacy header 612, thereby allowing a receiver to perform a secondary gain adjustment.

[0073] To get constant received power, the training symbols transmitted from the two (i.e. the first and second) antennas should be incoherent. This incoherence can be achieved by splitting the short symbols and the long symbols in the frequency domain. In other words, short symbols 704A use one half of the bins used by short symbols 601 and short symbols 704B use the other half of the bins (thus, 704A+704B=601). Similarly, long symbols 705A use one half of the bins used by long symbols 602 and long symbols 705B use the other half of the bins (thus, 705A+705B=602). In one embodiment, each antenna can transmit using both halves of the bins but at different times.

[0074] Because only half of the bins are used on one transmit antenna, the power per bin should be doubled for the split short symbols and long symbols. Notably, the received power will remain constant starting from the split short symbols. Therefore, the gain setting during the split short symbols will be valid for the data

symbols. At the receiver, the channel estimation is pulled out for half of the bins at a time and can be combined and smoothed after both halves are available.

[0075] The splitting can be done in various ways. For example, in one embodiment, long symbols 705A can use bins -26, -24, … -2, 1, 3, …25, whereas long symbols 705B can use bins -25, -23, … -1, 2, 4, … 26. In another embodiment, long symbols 705A can use bins -26, -24, …-2, 2, 4, …26 and long symbols 705B can use bins -25, -23, …-1, 1, 3, …25. Note that if the peak-to-average ratio (PAR) of long symbol 602 is 3.18 dB and the data in each bin remains the same after splitting, then the first bin use embodiment yields a PAR of 5.84 dB and 6.04 dB for long symbols 705A and 705B, respectively, whereas the second bin use embodiment yields a PAR of 5.58 dB and 5.85 dB for long symbols 705A and 705B, respectively.

[0076] Notably, split short and long symbols can be generalized to any plurality of data streams. For example, as shown in Figure 7B, if there are three streams, then the bins should be split into three groups, A, B, and C, interleaved and spaced evenly across all the bins. Thus, for the split shorts, the first antenna can transmit short symbols 414A (using bins A), the second antenna can transmit short symbols 414B (using bins B), and the third antenna can transmit short symbols 414C (using bins C).

[0077] For the split long symbols, the first antenna can transmit long symbols 415A, 415B, and 415C (using bins A, B, and C, respectively) sequentially, the second antenna can transmit long symbols 415B, 415C, and 415A (using bins B, C, and A, respectively) sequentially, and the third antenna can transmit long symbols 415C, 415A, and 415B (using bins C, A, and B, respectively) sequentially. This rotation pattern allows channel

estimation for all bins and ensures orthogonality in the frequency domain at all times. One exemplary long sequence for two steams at 20 MHz can be $L_{-26:26}$={ -1 1 -1 1 1 1 -1 -1 -1 -1 1 1 1 1 -1 1 -1 1 -1 -1 1 1 1 -1 1 1 0 -1 1 1 -1 -1 1 -1 -1 1 -1 -1 1 -1 -1 1 1 1 1 -1 1 1 1 1 1 1 1}, wherein long symbol 705A uses bins [-26:2:-2 2:2:26] with a PAR of 2.73 dB and long symbol 705B uses bins [-25:2:-1 1:2:25] with a PAR of 2.67 dB.

[0080] An exemplary sequence for 3 streams at 20 MHz can be $L_{-26:26}$={-1 -1 1 1 1 1 1 -1 -1 -1 1 -1 -1 -1 -1 -1 1 1 1 1 1 1 1 -1 -1 1 0 1 -1 -1 -1 1 -1 1 -1 1 -1 1 1 -1 1 -1 -1 1 1 -1 1 1 -1 1 -1 -1 1}, wherein the first tone set is [-26:3:-2 2:3:26] with a PAR of 3.37 dB, the second tone set is [-25:3:-1 3:3:24] with a PAR of 3.10 dB, and the third tone set is [-24:3:-3 1:3:25] with a PAR of 3.10 dB. An exemplary sequence for 4 streams at 20 MHz can be $L_{-26:26}$={-1 1 1 1 1 -1 -1 -1 1 -1 1 1 1 -1 1 1 -1 1 -1 -1 -1 1 1 1 -1 1 0 1 1 -1 1 -1 -1 1 -1 -1 -1 -1 1 -1 -1 -1 1 1 1 1 -1 1 1 -1 1 1 1}, wherein the first tone set is [-26:4:-2 3:4:23] with a PAR of 3.05 dB, the second tone set is [-25:4:-1 4:4:24] with a PAR of 3.05 dB, the third tone set is [-24:4:-4 1:4:25] with a PAR of 3.11 dB, and the fourth tone set is [-23:4:-3 2:4:26] with a PAR of 3.11 dB.

[0081] An exemplary long sequence for 1 stream at 40 MHz can be:

$$L_{-58,+58}=\{-1\ 1\ 1\ 1\ 1\ -1\ 1\ 1\ -1\ -1\ -1\ -1\ 1\ 1\ 1\ 1\ 1\ -1\ 1\ 1\ -1\ 1\ -1\ 1\ 1\ -1\ -1\ 1\ 1\ -1\ -1\ 1\ 1\ -1\ -1\ -1\ -1\ -1\ 1\ -1\ 1\ 1\ -1\ -1\ -1\ 1\ -1\ -1\ 1\ -1\ -1\ 1\ 1\ -1\ 1\ 1\ 1\ 0\ 0\ 0\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ -1\ 1\ 1\ 1\ -1\ 1\ -1\ 1\ 1\ -1\ -1\ 1\ -1\ -1\ 1\ -1\ 1\ -1\ 1\ -1\ -1\ -1\ -1\ 1\ 1\ -1\ 1\ -1\ -1\ -1\ 1\ -1\ 1\ -1\ 1\ -1\ 1\ 1\ -1\ 1\ -1\ -1\ 1\ -1\ 1\ 1\ 1\}$$

[0082] An exemplary long sequence for 2 streams at 40 MHz can be:

$L_{-58,+58}$ = {-1 1 1 1 1 -1 1 1 -1 -1 -1 -1 1 1 -1 1 1 -1 1 1 -1 1 -1 1 1 -1
-1 1 1 -1 -1 1 1 -1 -1 -1 1 -1 1 -1 1 1 1 -1 -1 1 -1 -1 -1 -1 -1 -1 1 -1 1 -1 1
0 0 0 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 -1 1 -1 1 1 -1 -1 1 -1 1 1 -1 1 -1
1 -1 -1 -1 -1 1 1 -1 1 -1 -1 -1 -1 -1 1 -1 1 1 1 1 -1 1 1 -1 1 -1 1 1 1}

wherein the first tone set is [-58:2:-2 2:2:58] and the second tone set is [-57:2:-3 3:2:57].

[0083] An exemplary long sequence for 3 streams at 40 MHz can be:

$L_{-58,+58}$={-1 -1 -1 -1 -1 -1 1 1 1 -1 -1 -1 -1 -1 -1 1 1 1 -1 -1 -1 -1 -1 -1 1 1 1 1
1 1 1 1 1 -1 -1 -1 -1 -1 -1 1 1 1 1 1 1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 0
0 0 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 -1 -1 -1 1 1 1 1 1 1 1 1 1 -1 -1
-1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 -1 -1 -1 }

wherein the first tone set is [-58:3:-4 2:3:56], the second tone set is [-57:3:-3 3:3:57], and the third tone set is [-56:3:-2 4:3:58].

[0084] An exemplary long sequence for 4 streams at 40 MHz can be:

$L_{-58,+58}$={-1 1 -1 -1 -1 1 1 -1 1 1 1 -1 1 1 1 1 -1 -1 -1 -1 1 -1 -1 -1 1 -1 1 -1
1 -1 -1 -1 -1 1 1 -1 -1 -1 1 1 1 1 -1 1 -1 -1 1 -1 1 1 1 1 1 -1 -1 -1 1 1 0
0 0 -1 1 1 -1 -1 -1 -1 -1 1 1 1 -1 1 1 -1 -1 -1 1 -1 1 -1 -1 -1 1 -1 1 -1
-1 1 1 1 1 1 1 1 -1 -1 -1 -1 1 1 -1 1 1 1 -1 1 1 -1 -1 1 -1 1 -1 -1 1 1}

[0085] where the first tone set is [-58:4:-2 5:4:57], the second tone set is [-57:4:-5 2:4:58], the third tone set is [-56:4:-4 3:4:55], and the fourth tone set is [-55:4:-3 4:4:56]. Note that to reduce PAR, a random search of the data pattern can be performed.

[0086] The short symbols can be split similarly, while recognizing that only 1 out of 4 bins is used. For example, in one embodiment, short symbols 704A can use bins -24, -16, -8, 4, 12, 20 whereas short symbols 704B can use bins -20, -12, -4, 8, 16, 24. In another embodiment, short symbols 704A can use bins -24, -16, -8, 8, 16, 24 whereas short symbols 704B can use bins -20, -

12, -4, 4, 12, 20. If the PAR of short symbol 601 is 2.09 dB, then the first bin use embodiment yields a 4.67 dB PAR for both short symbols 704A and 704B, whereas the second bin use embodiment yields a PAR of 4.32 dB for short symbol 704A and a PAR of 2.79 dB for short symbol 704B. Note that an exhaustive search for the first bin use embodiment yields a minimum PAR of 4.26 dB. A similar exhaustive search for the second bin use embodiment yields a minimum PAR of 1.68 dB for short symbol 704A with polarity {1 -1 1 -1 -1 -1}, and a minimum PAR of 2.79 dB for short symbol 704B with polarity {1 -1 -1 -1 -1 1}.

[0087] Note that there are only a small number of bins used for each split short. Therefore, when the channel is frequency selective and assuming that the split short symbols, the split long symbols and the SIGNAL symbols have the same transmitted power, the average received power can be significantly different in the split short symbols. This power differential can cause problems for receiver gain control. Therefore, in one embodiment, 24 bins can be used for the short symbols, thereby ensuring more bins for each split short symbol. In another embodiment, shifted shorts with 12 bins can be used in all data streams, but each data stream can use a different frequency shift, i.e. 1, 2, or 3 bins, from the original shorts. This frequency shift can ensure the continuity in received power from the shifted short symbols to the split long symbols and thereafter. However, note that in this scheme, the number of transmit data streams that can be supported is no more than four. Additionally, the period of the shifted short symbols is longer than that of the legacy short symbol, which may require modifications in frequency offset estimation implementations.

[0088] As noted above, the legacy header can be transmitted from a set of antennas. In the event that the set includes a plurality of antennas, beam-forming effects can occur. To achieve omni-directional transmission, the complex contribution of each frequency bin to each of the antennas can be weighted. For example, phase shifts (e.g. phase ramps or alternatively any type of phase shift) and/or phase magnitudes can be applied across the bins on the other antennas so that different bins experience different beam forming. An exemplary technique to create phase ramps includes cyclic delayed diversity (CDD), which is well known to those skilled in wireless technology.

[0089] An example of phase magnitudes would be to use the even bins on one antenna and the odd bins on another antenna. In another example of phase magnitudes, all positive frequency bins can be used for one antenna and all negative frequency bins can be used for another antenna. Thus, generally, weighting the contribution of each frequency bin independently to the two antennas can be used to create an omni-directional transmission.

[0090] Because a receiver capable of receiving MIMO packets should also be able to receive legacy 802.11a/g packets, a mechanism can be provided to label a MIMO packet different from a legacy packet. Additionally, if a packet is a MIMO packet, then the receiver also needs to know the number of transmitted data streams. In one embodiment shown in Figure 7C, a first set of bits in legacy SIGNAL symbol 603 can indicate a MIMO packet and a second set of bits in legacy SIGNAL symbol 603 can indicate the number of transmitted data streams. For example, a reserved bit R of SIGNAL symbol 603 can be set to "1" to indicate that a MIMO packet is being transmitted. Additionally, a predetermined number of

least significant bits in a length field 721 of SIGNAL symbol 603 can be used to indicate the number of transmitted data streams. Thus, if two least significant bits are used, then the length value in length field 721 would be rounded to the third least significant bit.

[0091] Note that after a MIMO receiver decodes legacy SIGNAL symbol 603, it can check the reserved bit R. If that bit is "0", then the packet is a legacy packet, and the length value in length field 721 is the true packet length in bytes. However, if the reserved bit is "1", then the packet is a MIMO packet, and the last two bits of the LENGTH field are the number of transmitted data streams. In the latter case, the length of the packet is accurate within 2 bytes. Notably, a legacy receiver only uses the length value to compute the time it should refrain from transmission. Therefore, the value in length field 721 need not be very accurate for a legacy device. Advantageously, and noted above, the true length of each data stream can be included in the MIMO SIGNAL symbols (e.g. 706A and 706B in Figure 7A). Therefore, a MIMO receiver can effectively ignore the value stored in length field 721.

[0092] In another embodiment, a 'modulo' operation can be used to represent the number of MIMO streams. Specifically, if the number of data bytes of the packet is L, the number of data bytes per symbol is B, and the number of service and tail bytes is C, then the number of symbols required is

$$N_{sym} = \left\lceil \frac{L+C}{B} \right\rceil$$

wherein $\lceil\ \rceil$ stands for rounding up to the nearest integer. Suppose the number of data streams is M, and $M \le B$. In this case, the modified length is

$$\widetilde{L} = B \cdot (N_{sym} - 1) - C + M$$

[0093] Note that the number of symbols computed by the legacy device is still N_{sym}:

$$\hat{N}_{sym} = \left\lceil \frac{\widetilde{L} + C}{B} \right\rceil = \left\lceil \frac{B \cdot (N_{sym} - 1) + M}{B} \right\rceil = N_{sym}$$

[0094] The MIMO device can compute the number of streams by:

$$\hat{M} = (\widetilde{L} + C) \bmod B = M$$

[0095] If $M = B$, then $\hat{M}$ will be zero. In this case, a mapping $\hat{M} = B$ will take place. Note that this same technique can be applied to signaling other information besides the number of streams, wherein the information to be signaled is encoded as M above.

[0096] In yet another embodiment, an encoding symbol 722 can be inserted after legacy SIGNAL symbol 603 to indicate the MIMO packet (and retaining the reserved bit for other uses). Encoding symbol 722 can include MIMO SIGNAL symbols with flipped pilot tones (i.e. +/-)(with respect to regular symbols that would otherwise appear in that location). For example, encoding symbol 722 can include modified SIGNAL symbols 706A' and 706B' with BPSK modulation for robustness. In this embodiment, the MIMO receiver can determine whether an incoming packet is a MIMO packet or legacy packet based on the phase of the pilots of encoding symbol 722. If it is a MIMO packet, then the number of transmitted data streams can be extracted and the rest of the packet can be detected in MIMO-compliant manner. Otherwise, the packet is treated as legacy 802.11a/g packet.

Pilots

[0097] Pilots are inserted in 802.11a/g systems for frequency offset and phase noise tracking. In a MIMO system where multiple radios are used at the transmitter and the receiver, different transmit and receive chains may experience common or independent phase noise. In accordance with one aspect of the invention, pilot tracking schemes can be advantageously provided for joint, per transmit chain, or per transmit-receive pair.

[0098] Due to frequency offset between the transmitter and receiver and phase noise, the phase of the received data symbols can vary during the transmission of the packet. To track and correct the phase variations, four pilot bins are inserted into each OFDM symbol in 802.11a/g in the format of $[1\ \ 1\ \ 1\ \ -1]*p_l$, where $[1\ \ 1\ \ 1\ \ -1]$ is the pattern across pilot bins and p_l is the pilot polarity for symbol *l*. For MIMO OFDM symbols, both the 4-bit pattern and the pilot polarity sequence can be generalized to multiple spatial streams.

[0099] In one embodiment, the same pilot format as 802.11a/g is duplicated in all transmitted data streams. For example, if the pilot polarity sequence for 802.11a/g symbols is $p_0, p_1, p_2, p_3, p_4 \cdots$, then the following pilot polarity can be used for the MIMO symbols, where different rows represent different transmitted streams:

$$\begin{array}{cccccc} p_0 & p_1 & p_2 & p_3 & p_4 & \cdots \\ p_0 & p_1 & p_2 & p_3 & p_4 & \cdots \\ p_0 & p_1 & p_2 & p_3 & p_4 & \cdots \end{array}$$

[00100] Because the pilot polarity is the same across the different spatial streams, a fixed beam-forming pattern in the pilot bins will result if the 4-bit pilot pattern is also duplicated across streams. To ensure that bad bins do not stay in a null all the time, phase

shifts can be added and cyclicly rotated across the 4 bins from symbol to symbol. For example:

	sym1	sym2	sym3	sym4
Ant 1	[1 1 1 1]	[1 1 1 1]	[1 1 1 1]	[1 1 1 1]
Ant 2	[1 j -1 -j]	[j -1 -j 1]	[-1 -j 1 j]	[-j 1 j -1]

[0100] The pilots on the second antenna (Ant 2) have 0, 90, 180, and 270 degrees of phase shift in the first symbol (sym1), and are cyclicly rotated to the left for subsequent symbols. Note that an initial pilot pattern of [1 1 1 1] for 4 pilot bins is used in the above example, but this scheme can be applied to any initial pilot pattern and more than 4 pilots bins. Thus, in general, pilots can be spaced across any frequency spectrum to perform estimations.

[0101] Pilot tracking can be performed in different ways. For example, if the phase noise is common across all transmit and receive chains (thereby allowing joint pilot tracking), then each receive chain can estimate the received signal in each pilot bin based on the channel estimation and the known pilot patterns. The complex conjugate of this estimate can then be multiplied to the actual received pilot signal. The results can be combined across pilot bins as well as chains. The phase of the final result is then the desired phase offset. In mathematical formulation, the received signal on receiver n in pilot bin k is represented by:

$$y_{n,k} = \sum_m H_{n,m,k} e^{j\theta} \cdot s_{m,k} + n_{n,k} \qquad \text{Eq. 1}$$

where $s_{m,k}$ is the pilot symbol of stream m, θ is the common phase offset, $H_{n,m,k}$ is the channel response, and

$n_{n,k}$ is the noise. The common phase offset is represented by:

$$\theta = \text{angle}(\sum_{n,k} y_{n,k} \cdot (\sum_{m} \hat{H}_{n,m,k} s_{m,k})^*) \qquad \text{Eq. 2}$$

where $\hat{H}_{n,m,k}$ is the estimated channel.

[0102] In contrast, if separate phase noise is present across different transmit chains (thereby necessitating pilot tracking per transmit chain), then MIMO detection algorithms can first be applied to the pilot bins to detect the pilots $\hat{s}_{m,k}$. Because $\hat{s}_{m,k} \approx s_{m,k} \cdot e^{j\theta_t(m)}$, where $\theta_t(m)$ is the phase offset of stream *m*, the phase difference between the decoded pilots and the ideal pilots can be averaged over the pilot bins of each data stream to generate the phase estimate,

$$\hat{\theta}_t(m) = \text{angle}(\sum_{k} \hat{s}_{m,k} \cdot s^*_{m,k}) \qquad \text{Eq. 3}$$

[0103] If phase noise is independent across transmit chains and receive chain (thereby necessitating pilot tracking per transmit-receive pair), the pilot polarity sequences can be modulated with orthogonal patterns so that the phase can be estimated separately for each transmit-receive pair. For example, if the number of transmitted data streams is, then the modulating pattern for stream *m*, wherein 1 ≤ *m* ≤ *M*, can be represented by $q_m(l) = e^{j\frac{2\pi}{M}(m-1)l}$, where $l \geq 0$ is the index of the MIMO symbols. For example, an exemplary modulated pilot polarity sequence for three transmitted streams could be:

$$\begin{array}{cccccc} p_0 & p_1 & p_2 & p_3 & p_4 & \cdots \\ p_0 & p_1 e^{j2\pi/3} & p_2 e^{-j2\pi/3} & p_3 & p_4 e^{j2\pi/3} & \cdots \\ p_0 & p_1 e^{-j2\pi/3} & p_2 e^{j2\pi/3} & p_3 & p_4 e^{-j2\pi/3} & \cdots \end{array}$$

[0104] Note that $\frac{1}{M}\sum_{l=k}^{k+M-1} q_m(l)q_n^*(l)=\delta_{m,n}$, where

$$\delta_{m,n}=\begin{cases}1 & for\ \ m=n\\ 0 & for\ \ m\neq n\end{cases}$$

i.e. the patterns are orthogonal across data streams over any interval of M symbols long with k representing a starting index of M orthogonal data symbols. In this case, the same 4-bit pilot pattern should be used for all streams to maintain orthogonality.

[0105] In one embodiment, the last $(M-1)$ symbols received on each antenna can be saved in a buffer. Thereafter, when a new symbol is received on each antenna, the complex conjugate of the pilots for these M symbols is multiplied to the symbols and summed, $\sum_l y_{n,k}(l)\cdot r_{m,k}^*(l)$, where $y_{n,k}(l)$ is the received signal on chain n in bin k for the l-th symbol, and $r_{m,k}(l)$ is the pilot symbol in stream m in bin k for the l-th symbol. The term $r_{m,k}(l)$ includes the bin pilot pattern, the original pilot polarity, and the orthogonal modulation. This computation is performed for all transmit-receive pairs (m,n) in all pilot bins k. The result can then be multiplied with the complex conjugate of the channel estimation, thereby yielding the orthogonally combined and channel corrected pilots, which can be represented by:

$$v_{n,m,k}=\sum_l y_{n,k}(l)\cdot r_{m,k}^*(l)\cdot H_{n,m,k}^* \qquad \text{Eq. 4}$$

[0106] The phase offset of stream m on receive antenna n is then estimated by averaging over the pilot bins, which is represented by:

$$\theta_{n,m}=\text{angle}(\sum_k v_{n,m,k})=\text{angle}(\sum_k\sum_l y_{n,k}(l)\cdot r_{m,k}^*(l)\cdot H_{n,m,k}^*) \qquad \text{Eq. 5}$$

[0107] For the first $(M-1)$ MIMO symbols, the joint or per transmit chain pilot tracking methods can be used since the history is not long enough. Moreover, any orthogonal pattern $q_m(l)$ (where m and l are defined as above) satisfying the following condition, can be used to modulate the pilot polarity sequence:

$$\frac{1}{M}\sum_{l=k}^{k+M-1} q_m(l)q_n^*(l) = \delta_{m,n}, \qquad \text{where} \quad \delta_{m,n} = \begin{cases} 1 & for\ \ m = n \\ 0 & for\ \ m \neq n \end{cases}$$

where m, n, l, k, and M are as defined above. For example, the modulated pilot polarity sequences for three transmitted streams of the above example would be:

$$\begin{matrix} p_0q_1(0) & p_1q_1(1) & p_2q_1(2) & p_3q_1(3) & p_4q_1(4) & \cdots \\ p_0q_2(0) & p_1q_2(1) & p_2q_2(2) & p_3q_2(3) & p_4q_2(4) & \cdots \\ p_0q_3(0) & p_1q_3(1) & p_2q_3(2) & p_3q_3(3) & p_4q_3(4) & \cdots \end{matrix}$$

[0108] Note that the inheritance of the 802.11a/g pilot polarity sequence is merely for similarity and can be abandoned all together, i.e. setting $p_l = 1$. This setting leads to a third and a fourth embodiment. In the third embodiment, the pilot polarity sequences are all ones, as represented by:

$$\begin{matrix} 1 & 1 & 1 & 1 & 1 & \cdots \\ 1 & 1 & 1 & 1 & 1 & \cdots \\ 1 & 1 & 1 & 1 & 1 & \cdots \end{matrix}$$

[0109] Similar to the first embodiment, cyclicly rotated phase shifts across pilot bins should be included to avoid fixed beam forming effect. Either joint pilot tracking or pilot tracking per transmit chain can be performed.

[0110] In the fourth embodiment, the pilot sequences are just $q_m(l)$, which can be represented by:

$$\begin{matrix} q_1(0) & q_1(1) & q_1(2) & q_1(3) & q_1(4) & \cdots \\ q_2(0) & q_2(1) & q_2(2) & q_2(3) & q_2(4) & \cdots \\ q_3(0) & q_3(1) & q_3(2) & q_3(3) & q_3(4) & \cdots \end{matrix}$$

[0111] In this case, the same pilot pattern across bins can be used for all streams. Pilot tracking per transmit-receive pair can be performed.

Data stream splitting

[0112] To form the MIMO SIGNAL symbols, the source data bits need to be split appropriately into multiple data streams. In 802.11a/g, convolutional codes of rate 1/2, 2/3 and 3/4 are used, and four modulation schemes are provided (i.e., BPSK, QPSK, 16QAM and 64QAM). The code rate and modulation scheme determine the number of bits in each OFDM symbol. For optimal MIMO performance, different modulations and coding rates should be allowed for different data streams. Therefore, the number of bits in each MIMO SIGNAL symbol can be different for the different data streams.

[0113] A typical coding block consists of an encoder and a puncturer, both of which are well known in the art of WLAN technology (e.g. puncturers are described in IEEE 802.11a, section 17.3.5.6). In accordance with one aspect of the invention, different codes can be constructed by using the same encoder but different puncturers, or by using the same puncturer but different encoders. If the same encoder is used, then splitting can be done either before the encoder or before the puncturer. On the other hand, if different encoders are used, then splitting must be done before the encoders. Splitting before the puncturer is referred to herein as a "shared" encoder, whereas splitting before the encoders

is referred to herein as "separate" encoders. Note that in 802.11a/g, both rate 2/3 and rate 3/4 codes are punctured from the rate 1/2 convolutional code. Therefore, either a shared encoder or separate encoders can be implemented.

[0114] Additional bits can be inserted before and after the source data bits to initialize and terminate the encoder. For example, in 802.11a/g, 16 service bits can be added before and 6 tail bits can be added after the source data bits. Therefore, in the case of separate encoders, these added bits can be inserted for each encoder.

[0115] Figure 8 illustrates an exemplary shared encoder system 800 for two spatial streams. In system 800, source data bits 801 can be provided to block 802, which adds the above-described service/tail bits. An encoder 803 receives the modified bits and generates n1 + n2 bits. A splitter 804 receives n1 + n2 bits and generates two spatial streams, which in turn are provided to puncturers 805A and 805B, respectively.

[0116] In one embodiment, for every n1 + n2 bits after encoder 803, the first spatial stream gets the first n1 bits, and the second spatial stream gets the last n2 bits. One exemplary chunk size is $n_i = N_{cbps}(i)$ (wherein N_{cbps} is the number of coded bits per symbol before puncturing), thereby splitting symbol by symbol. Another exemplary chunk size is $n_i = N_{cbps}(i)/\gcd(N_{cbps}(1), N_{cbps}(2))$ (wherein gcd() stands for the greatest common divider), thereby reducing the splitting chunk size while maintaining a proper ratio to reduce processing delay.

[0117] In this embodiment, the length fields in the MIMO SIGNAL symbols are set to the true length of the packet in bytes. The R14 fields in the MIMO SIGNAL

symbols are set to the individual data rates. The R14 field in the legacy SIGNAL symbol can be set to the data rate of the first data stream, or always set to the lowest data rate. As described in further detail below, the length field in the legacy SIGNAL symbol can be manipulated such that the number of symbols computed by a legacy device can be consistent with the actual duration of the packet.

[0118] Figure 9 illustrates another exemplary shared encoder system 900 for two spatial streams. In system 900, source data bits 901 can be provided to block 902, which adds the service/tail bits. An encoder 903 receives the modified bits and generates n1 + n2 bits. A puncturer 904 receives the n1 + n2 bits and generates an output code at a predetermined rate. A splitter 905 receives the n1 + n2 bits at the predetermined rate and generates the two spatial streams, i.e. n1 bits and n2 bits. Once again, for every n1 + n2 bits after encoder 903, the first spatial stream gets the first n1 bits, and the second spatial stream gets the last n2 bits.

[0119] Figure 10 illustrates an exemplary separate encoder system 1000 for two spatial streams. In system 1000, source data bytes 1001, i.e. N1 + N2, can be provided to a splitter 1002, which generates the two spatial streams (the first spatial stream getting the first N1 bytes and the second spatial stream getting the last N2 bytes). Blocks 1003A and 1003B receive and adds the service/tail bits to the N1 and N2 bytes, respectively. Encoders 904A and 904B receives the modified bytes, encode the modified bytes, and provide their encoded outputs to puncturers 905A and 905B, respectively.

[0120] Note that in a separate encoder system, the minimum data unit is in bytes because the length field in

the SIGNAL symbol is in bytes. Therefore, in this case, each stream can add 2 bytes of service bits at the beginning and 6 tail bits (~1 byte) at the end. Note that N1 can be the number of data bytes per symbol, or that divided by the greatest common divider of all numbers of data bytes per symbol. The number of bytes per symbol is integer for all data rates except for 9 Mbps, wherein each symbol contains 4.5 bytes. Therefore, in this case, the chunk size can alternate between 4 bytes and 5 bytes for data streams using 9 Mbps.

[0121] For example, assume there are two spatial streams (stream1 and stream2), and the number of data bytes per symbol is 27 (54 Mbps) and 4.5 (9 Mbps), respectively. Initially, the 2 service bytes can be sent to each of stream1 and stream2. Then, the first 25 (27-2) data bytes can be sent to stream1, the next 2 (4-2) data bytes can be sent to stream2, and the next 27 bytes to stream1, the next 5 bytes to stream2, and so on.

[0122] The values of the length field of the MIMO SIGNAL symbols are needed before the actual splitting is performed. The straightforward sequential splitting described above leads to a slightly complicated length calculation. Following is a simple length calculation that can be realized with slight modification to the splitting. First, the total number of symbols needed is computed:

$$N_{sym} = \left\lceil \frac{L+3M}{\sum_i B(i)} \right\rceil \qquad \text{(Eq. 6)}$$

wherein L is the total number of uncoded bytes in the packet, M is the number of data streams, and $B(i)$ is the number of uncoded bytes per symbol for stream i. $\lceil\ \rceil$ stands for rounding up to the nearest integer. If there are K

data streams using 9 Mbps and N_{sym} is odd, then recalculate:

$$N_{sym} = \left\lceil \frac{L + 3M + 0.5K}{\sum_i B(i)} \right\rceil \qquad \text{(Eq. 7)}$$

[0123] The number of bytes in the first data stream is then $L(1) = \lfloor B(1)N_{sym} - 3 \rfloor$, wherein $\lfloor \ \rfloor$ stands for rounding down to the nearest integer, the number of bytes in the second stream is $L(2) = \min(\lfloor B(2)N_{sym} - 3 \rfloor, L - L(1))$, and so on. A byte counter can be used for each stream. A stream can be skipped in the sequential splitting once its byte quota is met. Note that both equations 6 and 7 apply to general encoders and puncturers, a general number of data streams, and a general number of service/tail bits. Note that system 900 has only one length and therefore equations 6 and 7 do not apply to that system.

[0124] In 802.11a/g packets, the length field in the SIGNAL symbol is 12 bits long, which corresponds to a maximum packet size of 4095 bytes. In MIMO systems, packets larger than 4K bytes are desirable to maintain high payload efficiency. Therefore, signaling the exact packet length may require more bits than that can be contained in a single SIGNAL symbol.

[0125] In accordance with one embodiment, a pseudo-rate and pseudo-length can be used in the legacy SIGNAL symbol to indicate a rate and length that will occupy the same air-time as the MIMO packet. The lowest legal legacy rate (e.g. 6 Mbps for 802.11a/g) can be used to allow the packets to have the longest duration (4096 bytes @ 6 Mbps = 5.46 ms or for the longest valid 802.11 packet length, 2304 bytes @ 6 Mbps = 3.07 ms).

[0126] In one embodiment, for the MIMO SIGNAL symbols, a relative packet length can be used instead of the

absolute length to limit the number of required bits. The relative length is the total number of bytes that can be transmitted in a packet with the same number of symbols less the actual number of bytes transmitted, or roughly the number of padded bytes. As described above, the total number of bytes can be computed using the number of symbols in the packet (determined from the legacy SIGNAL symbol) and the data rates (encoded in the MIMO SIGNAL symbols).

[0127] For shared encoders, a single relative length is determined and transmitted only in the MIMO SIGNAL symbols of the first stream. The length fields in the other data streams can be reserved for other uses. For separate encoders, the relative length can be computed for each data stream and transmitted separately. Alternatively, a single relative length for all data streams can be computed and transmitted only in the first data stream (i.e. individual relative lengths can be derived from the total relative length for any byte allocation schemes agreed upon between the transmitter and receiver).

AGC and Channel Estimation

[0128] Referring back to Figure 7A, legacy short symbols 601 can be used for coarse frequency estimation, coarse timing estimation, and automatic gain control (AGC). Legacy long symbols 602 can be used for fine frequency estimation, fine timing estimation, and channel estimation. Legacy SIGNAL symbol 603 can include the information necessary to prevent legacy devices from stomping on the MIMO packet, as well as the signature of the MIMO packet and the number of data streams that are transmitted. Split short symbols 704A/B can be used for AGC for the MIMO section of the packet as well as for

antenna diversity selection (if appropriate). Split long symbols 705A/B can be used for the MIMO channel estimation. MIMO SIGNAL symbols 706A/B can include the length and modulation information of the transmitted data streams.

[0129] Because legacy header 612 may be transmitted from one antenna, while the MIMO header (including short symbols 704A/B, long symbols 705A/B, and SIGNAL symbols 706A/B) is transmitted from multiple antennas, the received power on each receive antenna may change from the legacy header to the MIMO header. In this case, split short symbols can be designed for the AGC to adjust the gain settings so that the input to ADC will be sized properly. Note that the AGC can use a single state machine for all receive chains, but each received chain may have a different gain applied to its corresponding received signal, depending on the signal size.

[0130] Additional timing recovery and frequency offset estimation, if necessary, are done jointly using the received split short symbols from the multiple antennas. This joint operation can be performed by combining the multiple received signals. The joint operation can also include choosing the best signal, and using that best signal for timing recovery and offset estimation.

[0131] In one embodiment, legacy header 612 can be transmitted from the best antenna to the intended receiver. This implies that the power increase from the legacy header to the MIMO header is no more than 10*log10(M) dB for a system with M spatial streams. The power increase may be higher for unintended receivers, but the average increase in dB is still 10*log10(M). Therefore, only fine gain changes are required.

[0132] The proposed split longs last 2M OFDM symbols, wherein M is the number of spatial streams. To compute

the channel estimation for any spatial stream, the corresponding bins used by each stream can be extracted from the FFT of the 2M OFDM symbols, averaged, and merged in the frequency domain. A smoothing filter can be applied to the frequency domain channel response to reduce estimation errors. For large M, the phase change across the 2M OFDM symbols can be significant. In one embodiment, the phase of each OFDM symbol can be corrected in time domain using the fine frequency estimation obtained from the legacy header before taking the FFT, averaging, and smoothing. Additional measurements of the phase changes (due to inaccuracy of the fine frequency estimate and phase noise) during the 2M long symbols can be used to properly align them prior to smoothing.

Detection Of MIMO Signals

[0133] A number of different techniques can be used to detect a MIMO signal. Two known techniques are the MMSE-LE and MMSE-DFE detection schemes. The Minimum Mean Square Error (MMSE) Linear Equalization (LE) or Decision Feedback Equalization (DFE) algorithms can be used to separate and detect the multiple data streams. For notation simplicity, only one sub-carrier is considered in the following description, wherein the same procedure would be repeated for each sub-carrier.

[0134] Suppose there are M transmitting antennas and N receiving antennas. If the frequency domain transmitted signal is x, the channel is H, the noise is n, and the received signal is y, where x is M-by-1, y and n are N-by-1, and H is N-by-M, then:

$$y = Hx + n, \qquad E(nn^*) = \sigma^2 I_N$$

[0135] It can be shown that the MMSE solution W that minimizes $E\|W^*y-x\|^2$ is:

$$W^*=(H^*H+\sigma^2 I_M)^{-1}H^*, \qquad R_e=\sigma^2(H^*H+\sigma^2 I_M)^{-1}$$

wherein R_e is the resulted error variance matrix. In the MMSE-LE algorithm, W^* is computed exactly as above and applied to y to detect all the data streams in parallel.

[0136] The MMSE-DFE detection algorithm performs successive cancellation using two steps (1) and (2). In step (1), nulling vectors can be computed. Computing the nulling vectors can in turn include three steps (a), (b), and (c). In step (a), the diagonal elements of R_e can be computed and the smallest element found. The smallest element corresponds to the transmit antenna that has the best signal quality. In step (b), the corresponding row of W^* can be computed. This will be the nulling vector of the selected transmit antenna. In step (c), the corresponding column in H can be deleted and M decremented by 1. Steps (a), (b), and (c) can be repeated until $M=0$.

[0137] In step 2, multiple data streams can be detected. Step 2, in turn, can include four steps (d), (e), (f), and (g). In step (d), y can be multiplied by the nulling vector for the best transmit antenna, thereby generating the raw decision of the best transmit antenna. In step (e), the corresponding column of H can be multiplied by this raw decision and the result subtracted from y. In step (f), steps (d) and (e) can be repeated for the next best transmit antennas until all antennas are decoded. In step (g)(which is optional), a decision directed feedback channel estimation update can be performed using the data decisions.

[0138] Note that the raw decision can be either a hard or a soft decision, wherein the hard decision is simply the constellation point closest to the channel corrected received symbol and the soft decision is a weighted sum of a few most likely constellation points (the weights being proportional to the likelihood of each constellation).

Viterbi bin weighting

[0139] In one embodiment, Viterbi decoders can be used to decode the convolutionally encoded data streams after detection at the receiver. For frequency selective fading channels, the reliability of the signal in different frequency bins can be different. Therefore, more weight can be assigned to data bits from good bins, and less weight can be assigned to data bits from bad bins in the Viterbi branch metrics computation. In one embodiment, the optimal bin weights can be proportional to the SNR (signal to noise ratio).

[0140] In 802.11a/g, where only one data stream is transmitted, SNR can be approximated by the square of the channel amplitude assuming noise is additive white Gaussian across the bins. However, in real systems, SNR often increases slower than the square of the channel amplitude due to channel estimation error, phase noise, and quantization noise. Therefore, in one embodiment, the channel amplitude can be used for bin weighting.

[0141] In MIMO systems, the detection SNR of each transmitted data streams can be computed assuming certain noise power density. Similarly, two different methods can be used to determine the Viterbi bin weights. In a first embodiment, the Viterbi bin weights can be proportional to the detection SNR. In a second

embodiment, the Viterbi bin weights can be proportional to the square root of the detection SNR.

[0142] For MMSE-DFE, the detection SNR computed from the MMSE formulas does not include the impact of error propagation, thereby leading to over-optimistic detection SNR for the data streams that are detected later. This over-optimistic detection SNR can undesirably degrade the decoder performance.

[0143] To improve the decoder performance, the impact of error propagation can be included in the noise term for SNR computation. Following is an example of computing the effective noise term for the second and third data streams:

$$\tilde{\sigma}_2^2 = \sigma_2^2 + \left|w_2^* h_1\right|^2 \cdot \sigma_1^2$$
$$\tilde{\sigma}_3^2 = \sigma_3^2 + \left|w_3^* h_2\right|^2 \cdot \tilde{\sigma}_2^2 + \left|w_3^* h_1\right|^2 \cdot \sigma_1^2$$

wherein σ_m^2 is the original noise term, w_m is the nulling vector, h_m is the channel, and $\tilde{\sigma}_m^2$ is the effective noise term for the *m*-th data stream.

Compensation For Different Noise Floors

[0144] The above derivation of the MMSE detector is based on the assumption that the noise power is the same across the components of *y*. This assumption is generally not true in real systems because of different noise floors and/or gain settings in the receiver chains. Therefore, the formula can be modified, as now described.

[0145] Assume the received signal at the antenna is

$$y = Hx + n, \qquad E\{nn^*) = \begin{bmatrix} \sigma_1^1 & & \\ & \ddots & \\ & & \sigma_N^1 \end{bmatrix}$$

[0146] After AGC, the received signal becomes

$$\tilde{y} = \begin{bmatrix} g_1 & & \\ & \ddots & \\ & & g_N \end{bmatrix} \cdot y = \tilde{H}x + \tilde{n}, \qquad \tilde{H} = \begin{bmatrix} g_1 & & \\ & \ddots & \\ & & g_N \end{bmatrix} \cdot H, \qquad E\{\tilde{n}\tilde{n}^*\} = \begin{bmatrix} g_1^2\sigma_1^2 & & \\ & \ddots & \\ & & g_N^2\sigma_N^2 \end{bmatrix}$$

wherein σ_n^2 is the noise floor, and g_n is the amplitude gain on the n-th receive antenna respectively. $\tilde{H}$ is the channel estimate since the channel is estimated after AGC.

[0147] In order to apply the MMSE solution, the noise variance should be scaled to the same value. To do that, let $K = \min_n(g_n\sigma_n)$ and define a scaling matrix

$$\Pi = \begin{bmatrix} \frac{K}{g_1\sigma_1} & & \\ & \ddots & \\ & & \frac{K}{g_N\sigma_N} \end{bmatrix}$$

[0148] The scaled channel is $H_{eq} = \Pi \cdot \tilde{H}$, and the resulting noise variance is $\sigma_{eq}^2 = K^2$, constant across all receive antennas. At this point, the nulling vectors W_{eq}^* can be computed using H_{eq} and σ_{eq}^2.

[0149] W_{eq}^* should be applied to $y_{eq} = \Pi \cdot \tilde{y}$. Instead of scaling $\tilde{y}$ for every symbol, it is preferable to compute $\tilde{W}^* = W_{eq}^* \cdot \Pi$ once and apply $\tilde{W}^*$ directly to $\tilde{y}$. For MMSE-DFE, the successive cancellation is done using $\tilde{y}$ and $\tilde{H}$. No scaling is necessary.

[0150] Figure 11 illustrates a portion of a receiver 1100 that can modify channel correction for a plurality of receiver chains. In receiver 1100, variable gain

amplifiers 1101 receive wireless signals (including associated channel information) from antennas and provide their amplified output to a channel inversion block 1102 for processing. An automatic gain control (AGC) block 1103 can generate AGC control values for variable gain amplifiers 1101. Channel inversion block 1102 can receive these AGC control values as well as a noise floor (also generated by AGC block 1103 to compute an appropriate channel correction. This channel correction can be provided to AGC 1103 to tune the AGC control values.

Compensation For Phase Errors

[0151] Due to phase noise, residual frequency offset, induced phase errors, and/or Doppler changes in the channel between the transmitter and receiver, the phase of the effective channel matrix H will change slowly throughout a packet. To model these effects, the effective channel can be written as $\Lambda_r \cdot H \cdot \Lambda_t$, wherein $\Lambda_r = \mathrm{diag}([\ e^{j\theta_r(1)} \quad e^{j\theta_r(2)} \quad \cdots \quad e^{j\theta_r(N)}\ ])$ and $\Lambda_t = \mathrm{diag}([\ e^{j\theta_t(1)} \quad e^{j\theta_t(2)} \quad \cdots \quad e^{j\theta_t(M)}\ ])$ capture the phase changes at the N receive antennas and the M transmit antennas respectively. The corresponding equalization matrix is $\Lambda_t^* \cdot W^* \cdot \Lambda_r^*$, which can be easily modified when the phase estimations are available.

[0152] As described above, different pilot schemes can be used that allow joint, per transmit chain, or per transmit-receive pair pilot tracking. For joint pilot tracking, only one common phase offset is estimated for all transmit and receive chains, i.e. Λ_t and Λ_r collapse into a scalar $e^{j\theta}$. Modification to the equalization matrix is simply multiplication by scalar $e^{-j\theta}$.

[0153] For pilot tracking per transmit chain, one phase estimation per transmitted data stream can be estimated, i.e. Λ_t and Λ_r collapse into one Λ_t. Therefore, the equalization matrix can be modified into $\Lambda_t^* \cdot W^*$. If needed, the phase estimates can be averaged over the transmit chains to get one common phase and applied as a scalar. The average can be derived from mean of the angles, $\theta = \frac{1}{M}\sum_m \theta_t(m)$, or angle of the mean (equivalently, sum), $\theta = \text{angle}(\sum_m e^{j\theta_t(m)})$.

[0154] For pilot tracking per transmit-receive pair, the orthogonally combined and channel corrected pilots are first derived (see equation 4). In a first embodiment, the phase offset of each element of the channel matrix *H*, $\theta_{n,m}(1 \le m \le M, 1 \le n \le N)$ is estimated from these pilots and converted into $\theta_t(m)(1 \le m \le M)$ and $\theta_r(n)(1 \le n \le N)$. Note the following mapping:

$$\begin{bmatrix} 1_N & & & & I_N \\ & 1_N & & & I_N \\ & & \ddots & & \vdots \\ & & & 1_N & I_N \end{bmatrix} \begin{bmatrix} \theta_t(1) \\ \theta_t(2) \\ \vdots \\ \theta_t(M) \\ \theta_r \end{bmatrix} = \begin{bmatrix} \theta_1 \\ \theta_2 \\ \vdots \\ \theta_M \end{bmatrix} \Leftrightarrow A \cdot \Theta_1 = \Theta_2 \qquad \Rightarrow \qquad \Theta_1 = pinv(A) \cdot \Theta_2$$

wherein 1_N is the *N*-by-1 vector of all 1's, I_N is the identity matrix of size *N*, $\theta_r = [\theta_r(1) \quad \theta_r(2) \quad \ldots \quad \theta_r(N)]^T$ is vector of the phase at the *N* receivers, and $\theta_m = [\theta_{1,m} \quad \theta_{2,m} \quad \ldots \quad \theta_{N,m}]^T$ is the phase vector of the *m*-th column of matrix *H*. The pseudo-inverse is the Least Square (LS) of the phase at the transmitter and receiver, which only depends on the number of transmit and receive antennas and therefore can be computed off-line.

[0155] Two implementation issues can be addressed here. First, the angles in Θ_2 should not be allowed to wrap around 2π from symbol to symbol, as a change of 2π in Θ_2 does not lead to a change of 2π in Θ_1. To unwrap Θ_2, the change in Θ_2 between the current symbol and the previous symbol is adjusted to within $(-\pi, \pi)$ by adding or subtracting 2π, and added to the previous Θ_2.

[0156] Second, if some of the angles in Θ_2 are unreliable (e.g. a weak component in the matrix channel), then the solution can also be unstable. The solution is to weight the components in Θ_2 according to their reliability when forming the cost function and solve a weighted-LS problem, i.e. minimizing $\|\Gamma(A \cdot \Theta_1 - \Theta_2)\|^2 = \|\Gamma A \cdot \Theta_1 - \Gamma\Theta_2\|^2$ instead of $\|A \cdot \Theta_1 - \Theta_2\|^2$, where Γ is a diagonal matrix with the weighting factors. The solution then becomes:

$$\Theta_1 = pinv(\Gamma A) \cdot \Gamma\Theta_2$$

[0157] Components with higher reliability should be weighted more, and lower reliability weighted less. One measurement of reliability is the magnitude of the channel components. The weights can be normalized by the maximum value, and if necessary, quantized to discrete levels for simplicity.

[0158] If needed, the estimated phase offsets of all transmit-receive antenna pairs can be averaged across the receive antennas to get one phase estimate per transmit antenna, $\theta_t(m) = \frac{1}{N}\sum_n \theta_{n,m}$ or averaged across all transmit and receive chains to get one common phase estimate $\theta = \frac{1}{MN}\sum_{n,m} \theta_{n,m}$.

[0159] In a second embodiment, θ_t and θ_r can be derived from combinations of the orthogonally combined and channel corrected pilots $v_{n,m,k}$ (see equation 4). The angle of the sum across pilot bins and receive antennas is the offset for each transmit antenna, $\theta_t(m) = \text{angle}(\sum_{n,k} v_{n,m,k})$. The angle of the sum across transmit antennas is the offset for each receive antenna, $\theta_r(n) = \text{angle}(\sum_{m,k} v_{n,m,k})$. The angle of the sum across all transmit and receive antennas, $\theta = \text{angle}(\sum_{n,m,k} v_{n,m,k})$, is computed and half of that is subtracted from both transmit and receive offsets to remove the bias, $\theta_t(m) = \theta_t(m) - \theta/2$ and $\theta_r(n) = \theta_r(n) - \theta/2$.

[0160] If needed, only the offset for each transmit chain is computed and applied, $\theta_t(m) = \text{angle}(\sum_{n,k} v_{n,m,k})$. Alternatively, only the common phase offset across all transmit and receive chains is computed and applied, $\theta = \text{angle}(\sum_{n,m,k} v_{n,m,k})$.

[0161] For continuous residual frequency offset correction, the common phase offset over all transmit and receive chains θ is used, because it captures the common phase shift due to residual frequency offset and suppresses fluctuations due to phase noise.

Closed Loop Transmit Optimization

[0162] If the MIMO transmitter has knowledge of the MIMO channel, it is possible to optimize the transmission scheme, including the number of data streams to transmit, the data rate to use for each stream, the sub-carriers to use for each stream, the selection of transmit antennas,

the transmit power for each antenna, and so on. Such optimization improves the robustness and the throughput of MIMO systems.

[0163] In a first embodiment and referring back to Figure 3, receiver 304 can assess the quality of the channel and feeds the information back to transmitter 301. The information can be either in the format of channel information (e.g. channel estimates or a detection pilot EVM) or in the format of a recommended transmit scheme. Note that a detection pilot EVM can be computed from the channel corrected pilots and known clean pilots, and therefore can be a good measurement of the signal quality. Two different packets can be used to send back the channel information: the CTS packets in the standard RTS/CTS exchange and the ACK packets.

[0164] In a second embodiment, transmitter 301 can estimate the channel using packets received from receiver 304. Reciprocity is assumed for this scheme where the same antennas are used on both sides for uplink and downlink. Therefore, transmitter 301 can determine the best transmit scheme based on the estimated channel.

[0165] Note that more data streams can be supported in channels with high spatial dimension, and fewer can be supported in channels with low dimension. The optimal number of data streams to use is determined based on the dimension of the MIMO channel estimate. For systems without transmit diversity, the same number of transmit antenna with the best channel are chosen from all available transmit antennas.

[0166] For systems with transmit diversity, each data stream can be phase shifted properly and transmitted from multiple antennas simultaneously to form a merged beam (called transmit beam forming (TxBF)). The BF procedure for each data stream can be performed using the

techniques described in U.S. Patent Application 10/682,381, entitled "Apparatus and Method of Multiple Antenna Transmitter Beamforming of High Data Rate…Signals", filed on October 8, 2003, and U.S. Patent Application 10/682,787, entitled, "Apparatus and Method of Multiple Antenna Receiver Combining of High Data Rate Wideband Signals", filed on October 8, 2003, both of which are incorporated by reference herein.

[0167] In general, different data streams are beam-formed toward the receive antennas to increase the receive SNR. This technique is particularly useful for the Access Point in systems with heavy downlink traffic. Transmit beam forming can be combined with high-rate MIMO by transmitting more than one but fewer than M unique data streams and using excess antennas to redundantly code and beam form the transmission in the intended direction.

[0168] In a discrete multi-tone (DMT) technique, the power and modulation type of each sub-carrier can be determined based on the channel estimates. Sub-carriers with good signal quality use more power and higher modulation levels than those with poor signal quality.

Receiver Selection Diversity

[0169] Due to cost and power consumption constraints, the number of receiver chains a MIMO receiver can have is usually limited. In contrast, the cost of RF antennas is much lower. It is therefore desirable to have more receive antennas than receiver chains and dynamically select the best receive antennas. This ability to dynamically select receive antennas yields diversity gain and improves the robustness of the system. To reduce the complexity and switching loss, in one embodiment, the RF antennas are divided into the same number of groups as

the number of receiver chains. Each antenna group is connected to its corresponding receiver chain through a switch.

[0170] In a first embodiment, fast antenna diversity can be used. In fast antennas diversity, each receiver chain can quickly sample the signal strength on the RF antennas that are connected to it, and select the antenna that has the strongest signal.

[0171] In a second embodiment, the selection criterion is based on the detection SNR. Channel estimation can be done for all RF antennas. For each possible combination of receive antennas, the detection SNR can be computed for all transmitted data streams. The minimum SNR can then be compared across all possible antenna combinations. The set of antennas that gives the maximum minimum SNR can be selected.

Rate Adaptation

[0172] Rate adaptation for MIMO systems is more challenging than for legacy 802.11a/g systems. Specifically, either implicit feedback (exploiting reciprocity) or explicit feedback (where explicit messages are used) is needed to assess the quality of the channel from each of the transmitted antennas.

[0173] This feedback can have varying levels of detail. At the coarsest level, a single acknowledgement can be used to indicate that all data in all streams is correct. This could make even a scheme with the same rate for all streams difficult, because it would be difficult to determine the number of transmit streams that could be supported as well as which ones are most optimal.

[0174] The next level of feedback would be to individually acknowledge each data stream. This

acknowledgement technique could allow for independent rate adaptation on each of the streams, though determining the optimal transmit antennas and the number of data streams supported could be difficult.

[0175] In yet another level of feedback, the intended receiver can perform channel measurements on the incoming packet during the channel estimation preambles and/or packet data portions. These estimates could determine the individual SNRs of each frequency/transmit antenna pair or could aggregate the information into bulk per-antenna values that could be used by the transmitter to adapt its rate. An intermediate solution is to just determine and report the SNR of a reduced set of frequency bins such as those of the pilot tone frequencies.

[0176] In one embodiment, the intended receiver can determine the best data rates that should be used based on its reception, and sends these data rates in the ACK to the transmitter. The transmitter decodes the ACK, retrieves the data rates, and applies these data rates to the next packet to that user. The information can include data rate information for each stream, or a single data rate and list of transmit antennas that can support it, or a single data rate and number of transmit antennas that can support it.

[0177] In another embodiment, the transmitter estimates the channel seen by the intended receiver from a special MIMO ACK sent by the intended receiver. To ensure reciprocity, the intended receiver will send the ACK using all the antennas it will receive with. This ACK only contains the legacy preamble and the MIMO preamble, without any data symbols. The transmitter can solicit for the special ACK when it needs to update the rate adaptation parameters or it could always use it.

(Note that the statistics of lost packets can be used as an auxiliary means to adapt the data rates slowly.)

[0178] The rate adaptation information, particularly if in the form of an explicit data rate or set of data rates, must be aged either over time or over failed transmissions to allow the data rate to drop in the event of multiple failures due to changing channel conditions.

Aggregation, Multiple Checksums and Partial ACKs

[0179] A MIMO header significantly increases packet overhead. On the other hand, to transmit the same number of bytes of information, a MIMO packet usually requires many fewer data symbols. Therefore, the overall efficiency of MIMO packets is much lower than the efficiency of legacy packets of the same size.

[0180] In one embodiment, to preserve the benefit of the high data rate provided by MIMO data transmission systems, only packets with size greater than a minimum threshold will be transmitted in MIMO formats. Packet aggregation can be used to increase the packet size, where several smaller data packets are aggregated into one large "super" packet.

[0181] In 802.11a/g, a CRC checksum is added at the end of the packet and passed to the physical layer. The receiver MAC checks for CRC errors in the output of the Viterbi decoder to decide if the packet has been received correctly. In high rate MIMO systems, the number of data bytes in a packet is usually much greater to improve efficiency as previously described. The error probability of these long packets is usually higher, and so is the cost of re-transmitting these packets.

[0182] To overcome this problem, multiple checksums can be included in every MIMO packet using one of two methods. Using a first method, individual checksums are

specified for each of the pre-aggregation packets. Using a second method, the post-aggregation super-packet can be divided into sections of equal length and checksums can be computed for each section and inserted after that section.

[0183] At the receiver (and after the decoder), the checksum can be examined for each packet/section to determine if that packet/section is received properly (e.g. using an acknowledgement bit-vector). If at least one packet/section is received correctly, then the receiver can send a partial ACK to the transmitter, indicating which of the packets/sections are received correctly. The transmitter will then only need to retry the failed packets/sections. To reduce MAC complexity, the MAC can choose to retransmit all sub-packets if any is in error. Note that the MAC could still use the individual bits in the acknowledgement bit-vector for rate adaptation purposes.

[0184] Although illustrative embodiments have been described in detail herein with reference to the accompanying figures, it is to be understood that the invention is not limited to those precise embodiments. They are not intended to be exhaustive or to limit the invention to the precise forms disclosed. As such, many modifications and variations will be apparent to practitioners skilled in this art.

[0185] For example, Figure 12 is a graph 1200 illustrating the data rate of several transmitter/receiver antenna configurations over relative distances. In graph 1200, line 1201 represents a 3 antenna transmitter and 3 antenna receiver configuration (3x3), line 1202 represents a 2x3 configuration, and line 1203 represents a 2x2 configuration. Note that a selected antenna configuration can be a compromise

between a peak data rate and robustness. Thus, in one embodiment, the 2x3 configuration represented by line 1202 may be chosen based on economic considerations.

[0186] Note that a "turbo" mode can be added to MIMO-SM as well as MIMO-AG. This turbo mode refers to wider channel bandwidths and is described in U.S. Patent Application 10/367,527, entitled "Receiving and Transmitting Signals Having Multiple Modulation Types Using Sequencing Interpolator", filed on February 14, 2003, as well as U.S. Patent Application 10/XXX,XXX, entitled "Multi-Channel Binding In Data Transmission", filed on November 6, 2003, both incorporated by reference herein. In general a turbo mode can be achieved by (1) double clocking or (2) channel bonding, i.e. using two regular 20 MHz channels (and potentially the gap in between them) together. Double clocking leads to the same sub-carrier structure as normal mode, but each sub-carrier is twice as wide. Channel bonding maintains the width of each sub-carrier but increases the number of sub-carriers. One specific example of channel bonding is to use 114 tones, from −58 to −2, and +2 to +58 (wherein 3 tones near DC, (-1, 0, +1), are not used).

[0187] Note that all of the embodiments of the above-described MIMO systems are applicable to turbo mode.

[0188] Figure 13 is a graph 1300 illustrating the data rate of several turbo and non-turbo antenna configurations over relative distances. As shown in graph 1300, turbo MIMO-SM delivers data rates up to 216 Mbps. However, turbo MIMO-AG can actually outperform MIMO-SM below 60 Mbps.

[0189] Accordingly, it is intended that the scope of the invention be defined by the following Claims and their equivalents.

CLAIMS

1. A time-division training pattern for multiple-input multiple-output packets, the pattern including:

a short symbol transmitted by a first antenna;

a first long symbol transmitted by the first antenna after transmission of the short symbol;

a legacy SIGNAL symbol transmitted by the first antenna after transmission of the first long symbol;

a second long symbol transmitted by a second antenna after transmission of the legacy SIGNAL symbol; and

SIGNAL symbols associated with multiple-input multiple-output data transmitted substantially simultaneously by the first antenna and the second antenna after transmission of the second long symbol.

2. A pattern for multiple-input multiple-output packets, the pattern including:

a short symbol transmitted by a first antenna and a second antenna, the short symbol being split between a predetermined set of short bins, wherein the first antenna is associated with a first set of short bins and the second antenna is associated with a second set of short bins;

a long symbol transmitted substantially simultaneously by the first antenna and the second antenna after transmission of the second short symbol, the long symbol associated with a first set of long bins and a second set of long bins, wherein the first antenna transmits using the first set of long bins before using the second set of long bins, and wherein the second antenna transmits using the second set of long bins before using the first set of long bins; and

SIGNAL symbols associated with multiple-input multiple-output data transmitted substantially simultaneously by the first antenna and the second antenna.

3. The pattern of Claim 2,

wherein the first set of short bins includes -24, -16, -8, 4, 12, 20, and

wherein the second set of short bins includes -20, -12, -4, 8, 16, 24.

4. The pattern of Claim 2,

wherein the first set of short bins includes -24, -16, -8, 8, 16, 24, and

wherein the second set of short bins includes -20, -12, -4, 4, 12, 20.

5. The pattern of Claim 2,

wherein the first set of long bins includes -26, -24, … -2, 1, 3, … 25, and

wherein the second set of long bins includes -25, -23, … -1, 2, 4, … 26.

6. The pattern of Claim 2,

wherein the first set of long bins includes -26, -24, … -2, 2, 4, … 26, and

wherein the second set of long bins includes -25, -23, … -1, 1, 3, … 25.

7. The pattern of Claim 2, further includes computing peak-to-average ratio (PAR) values for at least

two split patterns of the short bins and using the split pattern having a lowest PAR value.

8. The pattern of Claim 2, further includes computing peak-to-average ratio (PAR) values for at least two split patterns of the long bins and using the split pattern having an optimized PAR value.

9. The pattern of Claim 2, wherein the short symbol uses 24 bins.

10. The pattern of Claim 2, wherein the first set of short bins and the second set of short bins use a different frequency shift.

11. The pattern of Claim 10, wherein if the pattern is using 1 out of every N bins, then a frequency shift pattern can include 1 up to N-1 bins.

12. The pattern of Claim 2, wherein the short symbol uses 12 bins.

13. The pattern of Claim 2, further including a legacy header preceding the short and long symbols, the legacy header including:

a legacy short symbol transmitted by the first antenna;

a legacy long symbol transmitted by the first antenna after transmission of the legacy short symbol; and

a legacy SIGNAL symbol transmitted by the first antenna after transmission of the legacy long symbol.

14. The pattern of Claim 13, wherein the first antenna is implemented using a set of antennas.

15. The pattern of Claim 14, wherein complex weights are applied across bins of the first set of antennas, thereby mitigating beam-forming effects.

16. The pattern of Claim 15, wherein the complex weights include at least one of phase shifts or phase magnitudes, and wherein mitigating beam-forming effects creates a substantially omni-directional transmission.

17. The pattern of Claim 13, further including an encoding symbol transmitted after the legacy header to indicate that a multiple-input multiple-output packet is being transmitted.

18. The pattern of Claim 17, wherein the encoding symbol indicates at least a number of transmitted data streams.

19. The pattern of Claim 17, wherein the encoding symbol includes the SIGNAL symbols associated with multiple-input multiple-output (MIMO) data.

20. The pattern of Claim 19, wherein the encoding symbol includes flipped pilot tones, wherein the flipped pilot tones are different than those of a regular symbol that would appear in that location.

21. A method of sending a multiple-input multiple-output packet in a legacy device environment, the method comprising:

setting a reserved set of bits in a legacy SIGNAL symbol to a predetermined value, thereby indicating a multiple-input multiple-output signal is being transmitted.

22. A method of sending a multiple-input multiple-output (MIMO) packet in a legacy device environment, the method comprising:

using a set of bits in a legacy SIGNAL symbol to indicate information associated with the MIMO packet.

23. The method of Claim 22, wherein the set of bits includes a plurality of least significant bits of a length field of the legacy SIGNAL symbol.

24. The method of Claim 22, wherein the information includes a number of streams associated with the MIMO packet.

25. A method of sending a multiple-input multiple-output (MIMO) packet in a legacy device environment, the method comprising:

performing a 'modulo' operation on a set of bits in a legacy SIGNAL symbol to indicate information associated with the MIMO packet.

26. The method of Claim 25, wherein the information includes a number of streams.

27. A method of tracking and correcting phase variations of multiple received data symbols for a multiple-input multiple-output signal, the method comprising:

inserting a plurality of pilot bins into each data symbol.

28. The method of Claim 27, further including adding phase shifting using a pattern across the plurality of pilot bins.

29. The method of Claim 28, further including rotating the pattern of the phase shifting across the plurality of pilot bins.

30. The method of Claim 29, wherein rotating includes rotating the pattern cyclically across the plurality of pilot bins.

31. The method of Claim 28, wherein four pilot bins are inserted into each data symbol in a format of [1 1 1 -1]*p_l, wherein [1 1 1 -1] is a pattern across the four pilot bins and p_l is a pilot polarity for symbol *l*.

32. A method of tracking and correcting phase variations of multiple received data symbols for a multiple-input multiple-output (MIMO) signal, the method comprising:

providing orthogonal patterns across data streams over any interval of M data symbols long.

33. The method of Claim 32, wherein providing orthogonal patterns conforms to the equation:

$$\frac{1}{M}\sum_{l=k}^{k+M-1} q_m(l)q_n^*(l) = \delta_{mn} ,$$

wherein M represents a number of transmitted data streams, m represents a stream, k represents a starting

index of M orthogonal data symbols, l represented an index of MIMO symbols, and δ_{mn} is equal to 1 for m=n or equal to 0 for m≠n.

34. The method of Claim 33, wherein for M transmitted data streams, then a modulating pattern for stream m, wherein $1 \le m \le M$ and $l \ge 0$, is $q_m(l) = e^{j\frac{2\pi}{M}(m-1)l}$.

35. A method of joint pilot tracking across streams, the method comprising:

estimating a received signal in each pilot bin based on a channel estimation and known pilot patterns,

wherein a received signal on a receiver n in pilot K is represented by

$$y_{n,k} = \sum_m H_{n,m,k} e^{j\theta} \cdot s_{m,k} + n_{n,k}$$

wherein $s_{m,k}$ is a pilot symbol of stream *m*, θ is a common phase offset, $H_{n,m,k}$ is a channel response, and $n_{n,k}$ is noise,

wherein a common phase offset is represented by

$$\theta = \text{angle}\left(\sum_{n,k} y_{n,k} \cdot \left(\sum_m \hat{H}_{n,m,k} s_{m,k}\right)^*\right)$$

wherein $\hat{H}_{n,m,k}$ is the channel estimation.

36. A method of pilot tracking per transmit chain, the method including:

applying MIMO detection algorithms to pilot bins to detect the pilots $\hat{s}_{m,k}$, wherein $\hat{s}_{m,k} \approx s_{m,k} \cdot e^{j\theta_t(m)}$, where $\theta_t(m)$ is a phase offset of stream *m*; and

averaging a phase difference between decoded pilots and ideal pilots over the pilot bins of each data stream to generate a phase estimate $\hat{\theta}_t(m) = \text{angle}(\sum_k \hat{s}_{m,k} \cdot s^*_{m,k})$.

37. A method of pilot tracking per transmit/receive chains, the method including:

modulating pilot polarity sequences with orthogonal patterns, thereby estimating phase separately for each transmit/receive chain,

wherein if a number of transmitted data streams is M, then a modulating pattern for stream m, wherein 1 ≤ m ≤ M, can be represented by $q_m(l) = e^{j\frac{2\pi}{M}(m-1)l}$, where $l \geq 0$ is the index of the MIMO symbols.

38. The method of Claim 37, further including estimating a phase offset of stream m on a receive antenna n by averaging over a plurality of pilot bins, represented by

$$\theta_{n,m} = \text{angle}(\sum_k v_{n,m,k}) = \text{angle}(\sum_k \sum_l y_{n,k}(l) \cdot r^*_{m,k}(l) \cdot H^*_{n,m,k}) .$$

39. A method of splitting source data bits to form a multiple-input multiple-output signal, the method comprising:

adding bits to the source data bits to initialize and terminate an encoder, thereby creating modified source data bits;

providing the modified source data bits to the encoder, thereby creating encoded source data bits;

splitting the encoded source data bits into M data streams.

40. A method of splitting source data bits to form a multiple-input multiple-output signal, the method comprising:

splitting the source data bits into M data streams; and

adding bits to the M data streams to initialize and terminate M encoders, thereby creating M modified data streams.

41. The method of Claim 40, further including selecting a total number of bits such that when split across symbols for each of the M data streams, a number of symbols in each data stream is substantially equal.

42. A method of splitting source data bits to form a multiple-input multiple-output signal, the method comprising:

adding bits to source data bits to initialize and terminate an encoder, thereby creating modified source data bits;

providing the modified source data bits to the encoder, thereby creating encoded source data bits;

providing the encoded source data bits to a puncturer, thereby creating punctured source data bits; and

splitting the punctured source data bits into N data streams.

43. A method of indicating a length of a multiple-input multiple-output (MIMO) packet using a legacy SIGNAL symbol, the legacy SIGNAL symbol including a rate field and a length field, the length of the MIMO packet being longer than can be represented using the length field, the method comprising:

using the rate field and the length field to represent the length of the MIMO packet.

44. The method of Claim 43, wherein using the rate field and the length field includes providing a pseudo-rate value in the rate field and a pseudo-length value in the length field.

45. The method of Claim 44, wherein the pseudo-rate value is a lowest legacy rate and the pseudo-length value is an actual legacy length representing a transmit duration.

46. The method of Claim 43, wherein a MIMO SIGNAL symbol of the MIMO packet includes a relative packet length.

47. A pattern for multiple-input multiple-output (MIMO) packets, the pattern including:

a legacy header including a first plurality of short symbols for determining automatic gain control for receipt of the legacy header; and

a MIMO header including a second plurality of short symbols for facilitating automatic gain control for receipt of the MIMO header.

48. A pattern for multiple-input multiple-output (MIMO) packets, the pattern including:

a first short symbol transmitted by a plurality of antennas, the first short symbol being split between a predetermined set of short bins, wherein each of the plurality of antennas is associated with a subset of the short bins,

wherein the first short symbol is used for automatic gain control for a MIMO packet, the MIMO packet including the first short symbol.

49. The pattern of Claim 48, wherein the MIMO packet further includes a first long symbol, the pattern further including:

the first long symbol transmitted substantially simultaneously by the plurality of antennas, the first long symbol associated with sets of long bins, wherein each antenna transmits using a different order of the sets of long bins,

wherein the first long symbol is used for MIMO channel estimation.

50. The pattern of Claim 49,

wherein the plurality of antennas includes a first antenna and a second antenna,

wherein the first short symbol is transmitted by the first antenna and the second antenna after transmission of the legacy SIGNAL symbol, the first antenna being associated with a first set of short bins and the second antenna being associated with a second set of short bins;

wherein the first long symbol is transmitted substantially simultaneously by the first antenna and the second antenna, the first long symbol being associated with a first set of long bins and a second set of long bins, wherein the first antenna transmits using the first set of long bins before using the second set of long bins, and wherein the second antenna transmits using the second set of long bins before using the first set of long bins.

51. The pattern of Claim 49, further including:

SIGNAL symbols associated with MIMO transmitted substantially simultaneously by the first antenna and the second antenna after the first short symbol and the first long symbol.

52. The pattern of Claim 49, further including:

a second short symbol;

a second long symbol;

a legacy SIGNAL symbol,

wherein the second short symbol is used for automatic gain control for a legacy header, wherein the legacy header includes the second short symbol, the second long symbol, and the legacy SIGNAL symbol, and wherein the legacy header is transmitted before the MIMO header.

53. A pattern for multiple-input multiple-output (MIMO) packets, the pattern including:

a legacy header including a first plurality of long symbols used for legacy device channel estimation; and

a MIMO header including a second plurality of long symbols used for MIMO device channel estimation.

54. A pattern for multiple-input multiple-output (MIMO) packets, the pattern including:

a first long symbol transmitted by a plurality of antennas, the first long symbol being transmitted substantially simultaneously by the plurality of antennas, the first long symbol associated with sets of long bins, wherein each antenna transmits using a different order of the sets of long bins,

wherein the first long symbol is used for MIMO channel estimation for a MIMO packet, the MIMO packet including the first long symbol.

55. The pattern of Claim 54, wherein a long sequence for 2 streams at 20 MHz is $L_{-26:26}$={-1 1 -1 1 1 1 -1 -1 -1 -1 1 1 1 1 -1 1 -1 1 -1 -1 1 1 1 -1 1 1 0 -1 1 1 -1 -1 1 -1 -1 1 -1 -1 1 -1 -1 1 1 1 1 -1 1 1 1 1 1 1 1}, wherein a first tone set is [-26:2:-2 2:2:26] 73 dB and a second tone set is [-25:2:-1 1:2:25].

56. The pattern of Claim 54, wherein a long sequence for 3 streams at 20 MHz is $L_{-26:26}$={-1 -1 1 1 1 1 1 -1 -1 -1 1 -1 -1 -1 -1 -1 1 1 1 1 1 1 1 -1 -1 1 0 1 -1 -1 -1 1 -1 1 -1 1 -1 1 1 -1 1 -1 -1 1 1 -1 1 1 -1 1 -1 -1 1}, wherein a first tone set is [-26:3:-2 2:3:26], a second tone set is [-25:3:-1 3:3:24], and a third tone set is [-24:3:-3 1:3:25].

57. The pattern of Claim 54, wherein a long sequence for 4 streams at 20 MHz is $L_{-26:26}$={-1 1 1 1 1 -1 -1 -1 1 -1 1 1 1 -1 1 1 -1 1 -1 -1 -1 1 1 1 -1 1 0 1 1 20: -1 1 -1 -1 1 -1 -1 -1 -1 1 -1 -1 -1 1 1 1 1 -1 1 1 -1 1 1 1}, wherein a first tone set is [-26:4:-2 3:4:23], a second tone set is [-25:4:-1 4:4:24], a third tone set is [-24:4:-4 1:4:25], and a fourth tone set is [-23:4:-3 2:4:26].

58. The pattern of Claim 54, wherein a long sequence for 1 stream at 40 MHz is

$L_{-58,+58}$ ={-1 1 1 1 1 -1 1 1 -1 -1 -1 -1 1 1 1 1 1 -1 1 1 -1 1 -1 1 1 -1 -1 1 1 -1 -1 1 1 -1 -1 -1 -1 -1 1 -1 1 1 -1 -1 -1 1 -1 -1 1 -1 -1 1 1 -1 1 1 1 0 0 0 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 -1 1 -1 1 1 -1 -1 1 -1 -1 1 -1 1 -1 1 -1 -1 -1 -1 1 1 -1 1 -1 -1 -1 1 -1 1 -1 1 -1 1 1 -1 1 -1 -1 1 -1 1 1 1}

59. The pattern of Claim 54, wherein a long sequence for 2 streams at 40 MHz is

$L_{-58,+58}$ = {-1 1 1 1 1 -1 1 1 -1 -1 -1 -1 1 1 -1 1 1 -1 1 1 -1 1 -1 1 1 -1

-1 1 1 -1 -1 1 1 -1 -1 -1 1 -1 1 -1 1 1 1 -1 -1 1 -1 -1 -1 -1 -1 -1 1 -1 1 -1 1

0 0 0 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 -1 1 -1 1 1 -1 -1 1 -1 1 1 -1 1 -1

1 -1 -1 -1 -1 1 1 -1 1 -1 -1 -1 -1 -1 1 -1 1 1 1 1 -1 1 1 -1 1 -1 1 1 1}

wherein a first tone set is [-58:2:-2 2:2:58] and a second tone set is [-57:2:-3 3:2:57].

60. The pattern of Claim 54, wherein a long sequence for 3 streams at 40 MHz is

$L_{-58,+58}$={-1 -1 -1 -1 -1 -1 1 1 1 -1 -1 -1 -1 -1 -1 1 1 1 -1 -1 -1 -1 -1 -1 1 1 1 1

1 1 1 1 -1 -1 -1 -1 -1 -1 1 1 1 1 1 1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 0

0 0 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 -1 -1 -1 1 1 1 1 1 1 1 1 1 -1 -1

-1 -1 -1 -1 -1 -1 -1 -1 -1 -1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 1 -1 -1 -1 }

wherein a first tone set is [-58:3:-4 2:3:56], a second tone set is [-57:3:-3 3:3:57], and a third tone set is [-56:3:-2 4:3:58].

61. The pattern of Claim 54, wherein a sequence for 4 streams at 40 MHz is

$L_{-58,+58}$={-1 1 -1 -1 -1 1 1 -1 1 1 1 -1 1 1 1 1 -1 -1 -1 -1 1 -1 -1 -1 1 -1 1 -1

1 -1 -1 -1 -1 1 1 -1 -1 -1 1 1 1 1 -1 1 -1 -1 1 -1 1 1 1 1 1 -1 -1 -1 1 0

0 0 -1 1 1 -1 -1 -1 -1 -1 1 1 1 -1 1 1 -1 -1 -1 1 -1 1 -1 -1 -1 1 -1 1 -1

-1 1 1 1 1 1 1 1 -1 -1 -1 -1 1 1 -1 1 1 1 -1 1 1 -1 -1 1 -1 1 -1 -1 1 1}

where a first tone set is [-58:4:-2 5:4:57], a second tone set is [-57:4:-5 2:4:58], a third tone set is [-56:4:-4 3:4:55], and a fourth tone set is [-55:4:-3 4:4:56].

62. The pattern of Claim 54, wherein the MIMO packet further includes a first short symbol, the pattern further including:

the first short symbol being transmitted by the plurality of antennas, the first short symbol being split

between a predetermined set of short bins, wherein each of the plurality of antennas is associated with a subset of the short bins,

wherein the first short symbol is used for automatic gain control for the MIMO packet, the MIMO packet including the first short symbol.

63. The pattern of Claim 62,

wherein the plurality of antennas includes a first antenna and a second antenna,

wherein the first short symbol is transmitted by the first antenna and the second antenna after transmission of the legacy SIGNAL symbol, the first antenna being associated with a first set of short bins and the second antenna being associated with a second set of short bins;

wherein the first long symbol is transmitted substantially simultaneously by the first antenna and the second antenna, the first long symbol being associated with a first set of long bins and a second set of long bins, wherein the first antenna transmits using the first set of long bins before using the second set of long bins, and wherein the second antenna transmits using the second set of long bins before using the first set of long bins.

64. The pattern or Claim 63, further including:

SIGNAL symbols associated with MIMO transmitted substantially simultaneously by the first antenna and the second antenna after the first short symbol and the first long symbol.

65. A method of decoding a plurality of encoded data streams for a multiple-input multiple-output (MIMO) transmission, the method comprising:

for decoding, weighting data bits from good bins more heavily than bad bins.

66. The method of Claim 65, wherein the bin weights are proportional to a signal to noise ratio (SNR).

67. The method of Claim 65, wherein the bin weights are proportional to a square root of a signal to noise ratio (SNR).

68. The method of Claim 67, wherein weighting influences Viterbi branch metrics computation.

69. The method of Claim 68, wherein the bin weights are proportional to a signal to noise ratio (SNR).

70. The method of Claim 68, wherein the bin weights are proportional to a square root of a signal to noise ratio (SNR).

71. The method of Claim 65, further including determining the impact of error propagation based on the following equations for computing effective noise terms for second and third streams:

$$\tilde{\sigma}_2^2 = \sigma_2^2 + \left|w_2^* h_1\right|^2 \cdot \sigma_1^2$$
$$\tilde{\sigma}_3^2 = \sigma_3^2 + \left|w_3^* h_2\right|^2 \cdot \tilde{\sigma}_2^2 + \left|w_3^* h_1\right|^2 \cdot \sigma_1^2$$

wherein σ_m^2 is an original noise term, w_m is a nulling vector, h_m is a channel, and $\tilde{\sigma}_m^2$ is an effective noise term for an m-th data stream.

72. A method for modifying channel correction for a plurality of receiver chains, the method comprising:

receiving channel estimates for the plurality of receiver chains;

computing gain adjustment values for the plurality of receiver chains based on a noise floor and automatic gain control values; and

applying the gain adjustment values to the plurality of receiver chains.

73. A method of using phase estimates for a multiple-in multiple-out (MIMO) system, the method comprising:

using a single joint phase estimate from a plurality of data streams to compute a phase correction applicable to all data streams.

74. The method of Claim 73, wherein the plurality of data streams includes all data streams.

75. A method of providing phase estimations for each transmit/receive pair, the method including:

estimating a phase offset of each element of a channel matrix *H*, $\theta_{n,m}(1 \le m \le M, 1 \le n \le N)$, from pilots and converted the phase offset into $\theta_t(m)(1 \le m \le M)$ and $\theta_r(n)(1 \le n \le N)$.

76. The method of Claim 75, wherein

$$\begin{bmatrix} 1_N & & & & I_N \\ & 1_N & & & I_N \\ & & \ddots & & \vdots \\ & & & 1_N & I_N \end{bmatrix} \begin{bmatrix} \theta_t(1) \\ \theta_t(2) \\ \vdots \\ \theta_t(M) \\ \theta_r \end{bmatrix} = \begin{bmatrix} \theta_1 \\ \theta_2 \\ \vdots \\ \theta_M \end{bmatrix} \Leftrightarrow A \cdot \Theta_1 = \Theta_2 \quad \Rightarrow \quad \Theta_1 = pinv(A) \cdot \Theta_2$$

wherein 1_N is an *N*-by-1 vector of all 1's, I_N is an identity matrix of size *N*, $\theta_r = [\theta_r(1) \quad \theta_r(2) \quad \ldots \quad \theta_r(N)]^T$ is a phase vector at *N* receivers, and $\theta_m = [\theta_{1,m} \quad \theta_{2,m} \quad \ldots \quad \theta_{N,m}]^T$ is a phase vector of an *m*-th column of matrix *H*.

77. A method of optimizing transmission of a multiple-in multiple-out (MIMO) signal, the method comprising:

assessing a quality of a channel using a packet received by an intended receiver from a transmitter of the MIMO signal; and

sending a packet from the intended receiver to the transmitter, the packet including feedback information for optimizing transmission, the feedback information derived from a plurality of data streams previously transmitted substantially simultaneously.

78. The method of Claim 77, wherein the packet includes a CTS packet.

79. The method of Claim 77, wherein the packet includes an ACK packet.

80. The method of Claim 77, wherein the feedback information includes one of (1) channel estimates and (2) a detection pilot EVM computed from channel corrected pilots and known clean pilots.

81. The method of Claim 77, wherein the feedback information includes a data rate to be used by the transmitter.

82. The method of Claim 77, wherein the feedback information includes an indicator for at least one of a minimum data rate, a maximum data rate, a higher data rate, and a lower data rate to be used by the transmitter.

83. A method of optimizing transmission of a transmitted multiple-in multiple-out (MIMO) signal, the method comprising:

assessing a quality of a channel using a MIMO packet received by a transmitter for the MIMO signal from an intended receiver; and

determining optimized transmit information based on the MIMO packet.

84. A method of determining receiver selection for a multiple-input multiple-output (MIMO) signal in a diversity antenna system, wherein at least one receiver chain is connectable to a plurality of receive antennas, the method comprising:

for each receiver chain, selecting a receive antenna having a strongest signal.

85. A method of determining receiver selection for a multiple-input multiple-output (MIMO) signal in a diversity antenna system, wherein at least one receiver chain is connectable to a plurality of receive antennas, the method comprising:

determining possible combinations of receive antennas;

computing a signal to noise (SNR) for each combination; and

selecting a combination having a minimum SNR.

86. A method of selecting a split sequence comprising:

calculating power-to-average ratios (PARs) for a plurality of split sequences; and

selecting the split sequence having an optimized PAR.

87. The method of Claim 86, wherein the optimized PAR for short symbols is a lowest PAR provided by calculating the PAR for each of the plurality of split sequences.

88. The method of Claim 86, wherein the optimized PAR for long symbols is a relatively low PAR provided by calculating the PAR for each of a random set of the plurality of split sequences.

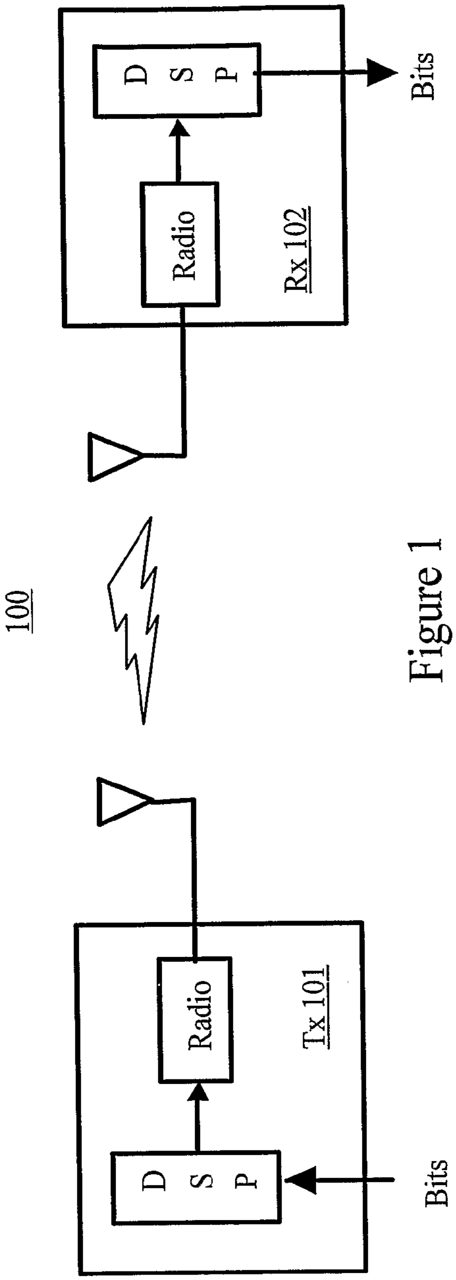

Figure 1

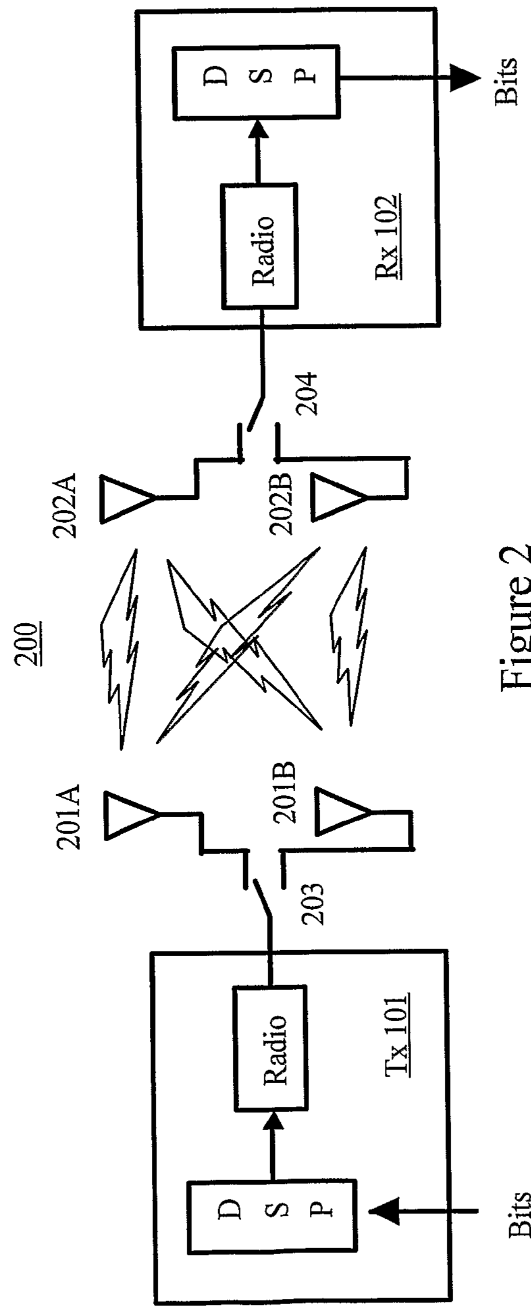

Figure 2

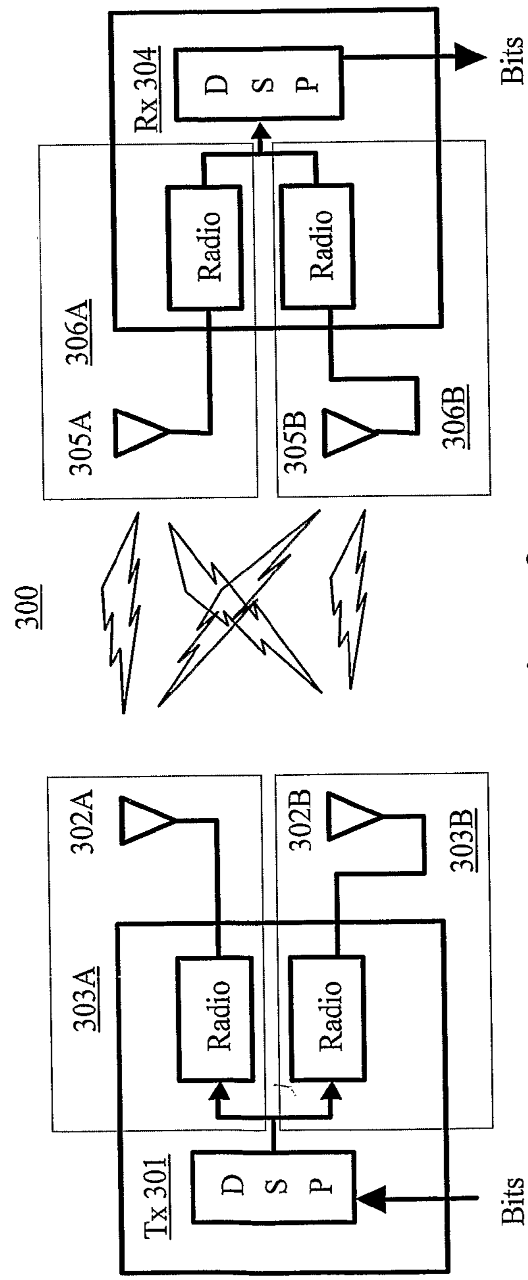
300
Tx 301
D S P
Bits
Radio
Radio
303A
303B
302A
302B
305A
305B
306A
306B
Radio
Radio
Rx 304
D S P
Bits

Figure 3

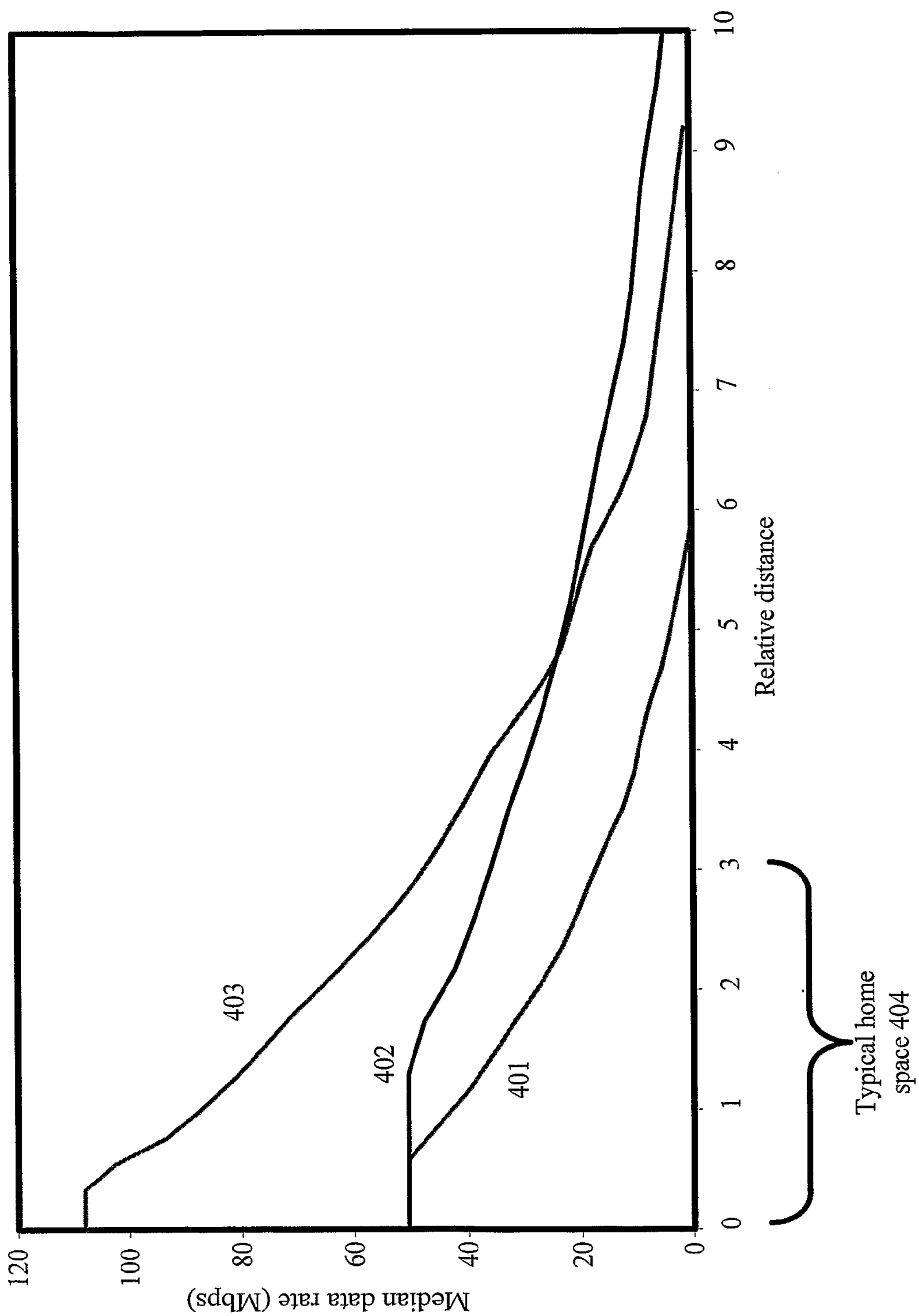
Median data rate (Mbps)
120
100
80
60
40
20
0
Relative distance
0
1
2
3
4
5
6
7
8
9
10
401
402
403
Typical home space 404

Figure 4

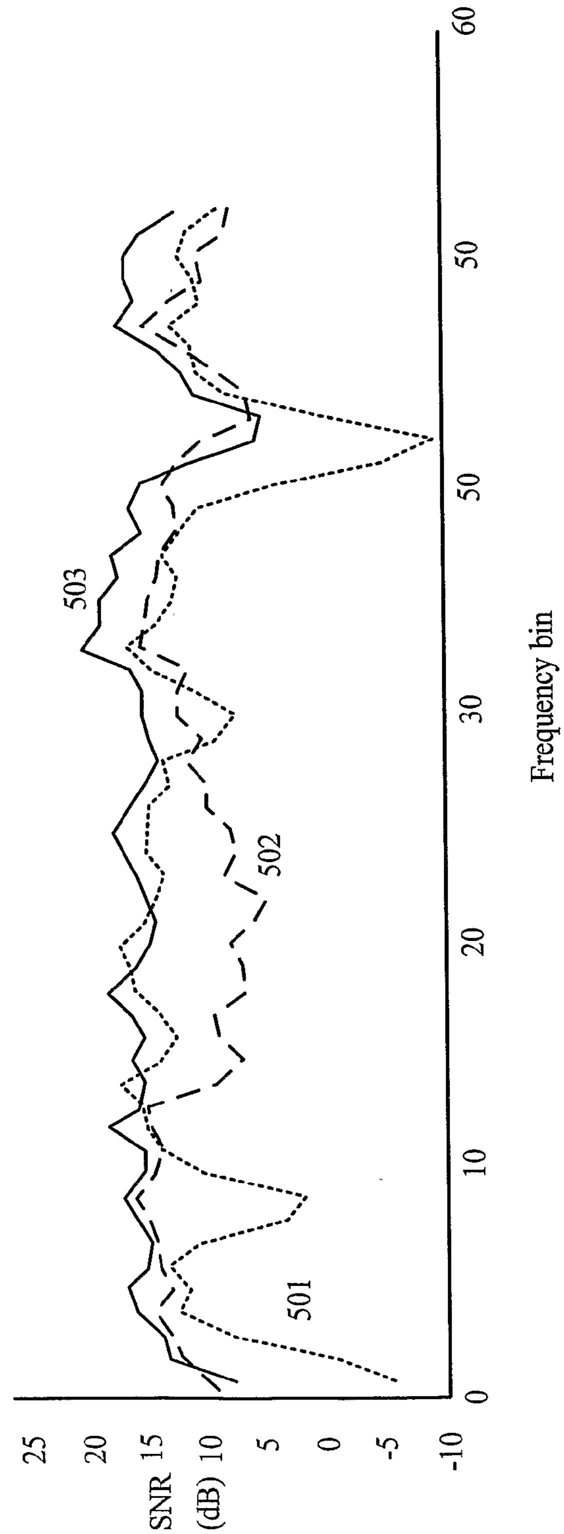
SNR
(dB)
25
20
15
10
5
0
-5
-10
0
10
20
30
50
50
60
501
502
503
Frequency bin

Figure 5

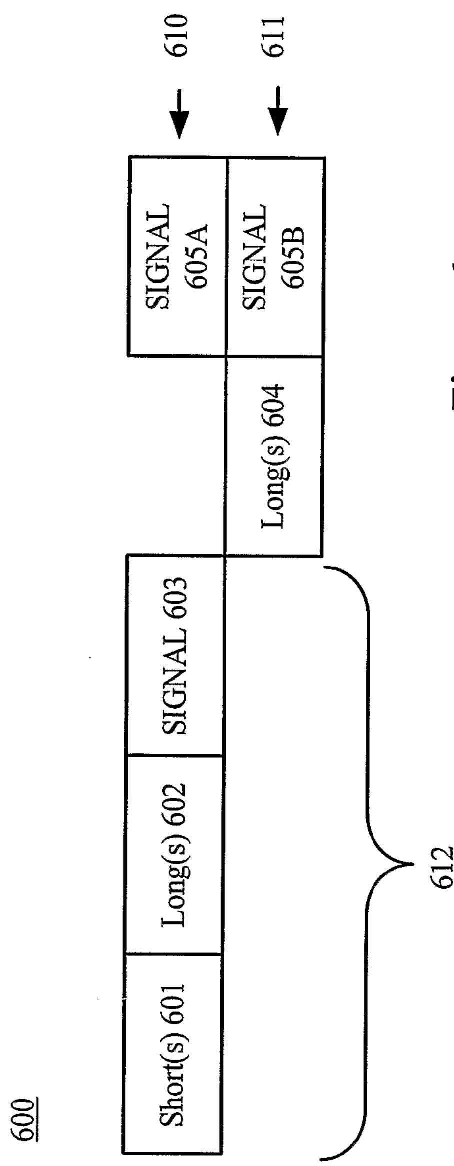

Figure 6

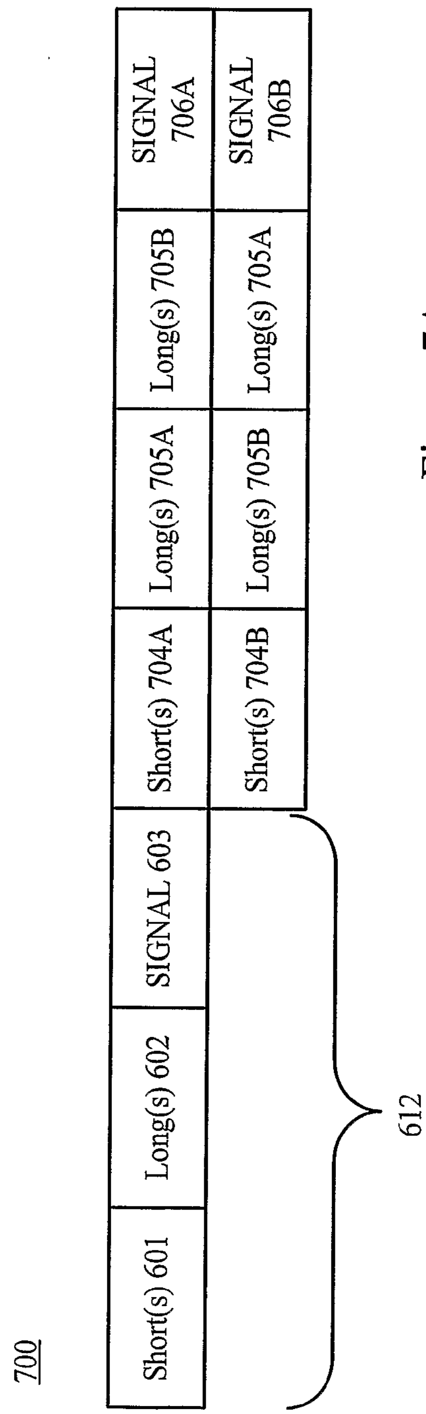

Figure 7A

710

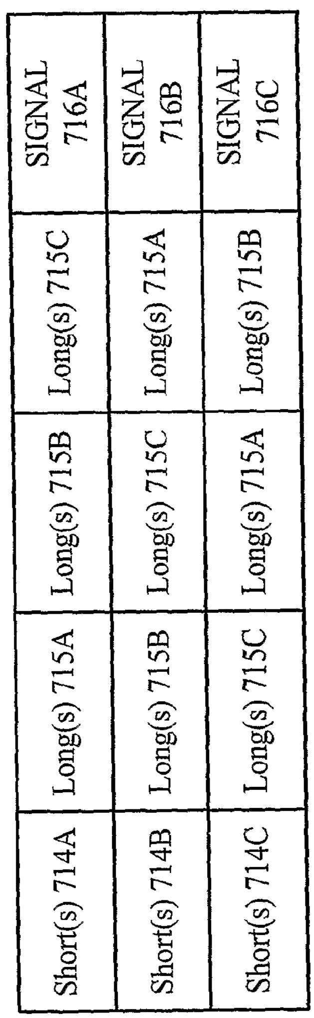

Figure 7B

720

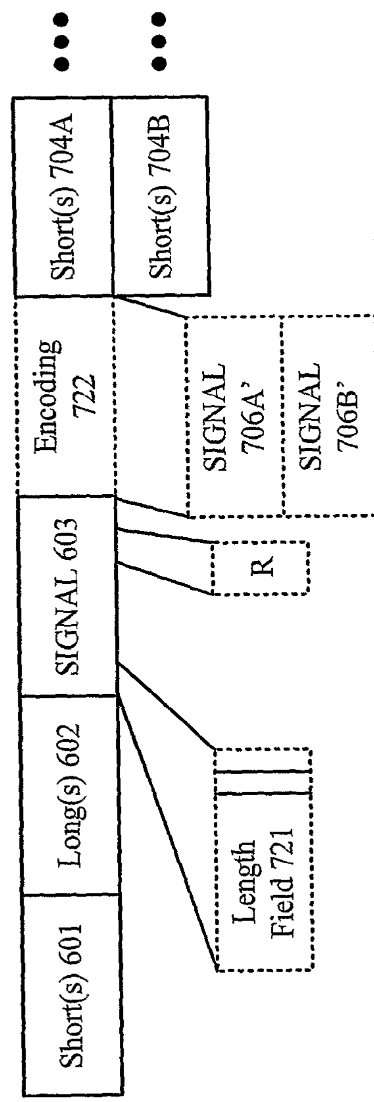

Figure 7C

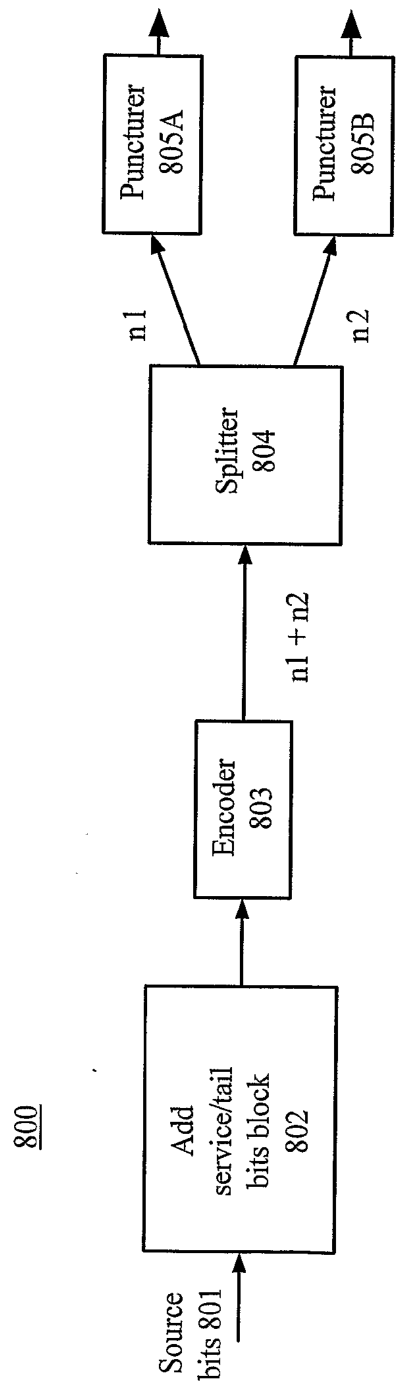
800
Source bits 801
Add service/tail bits block 802
Encoder 803
n1 + n2
Splitter 804
n1
n2
Puncturer 805A
Puncturer 805B

Figure 8

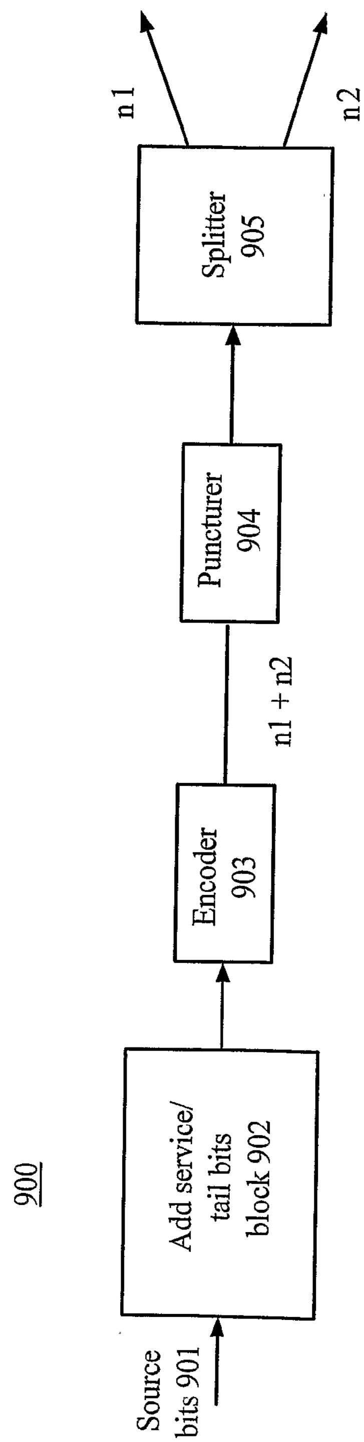
900
Source bits 901
Add service/ tail bits block 902
Encoder 903
n1 + n2
Puncturer 904
Splitter 905
n 1
n2

Figure 9

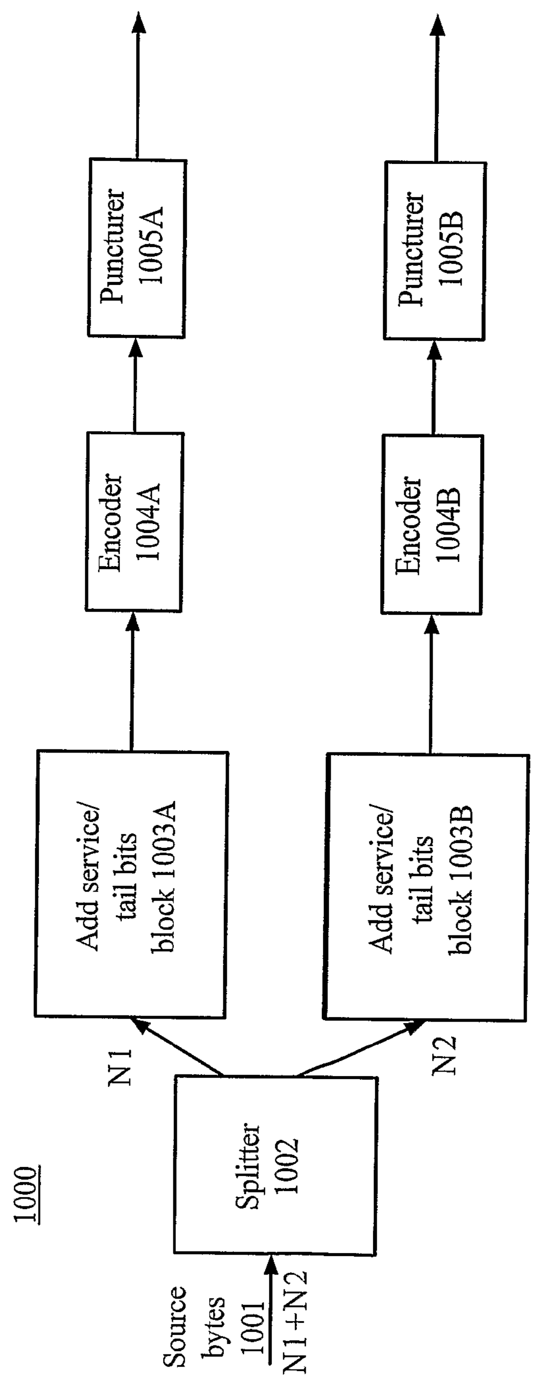
1000
Source bytes 1001
N1 +N2
Splitter 1002
N1
N2
Add service/ tail bits block 1003A
Add service/ tail bits block 1003B
Encoder 1004A
Encoder 1004B
Puncturer 1005A
Puncturer 1005B

Figure 10

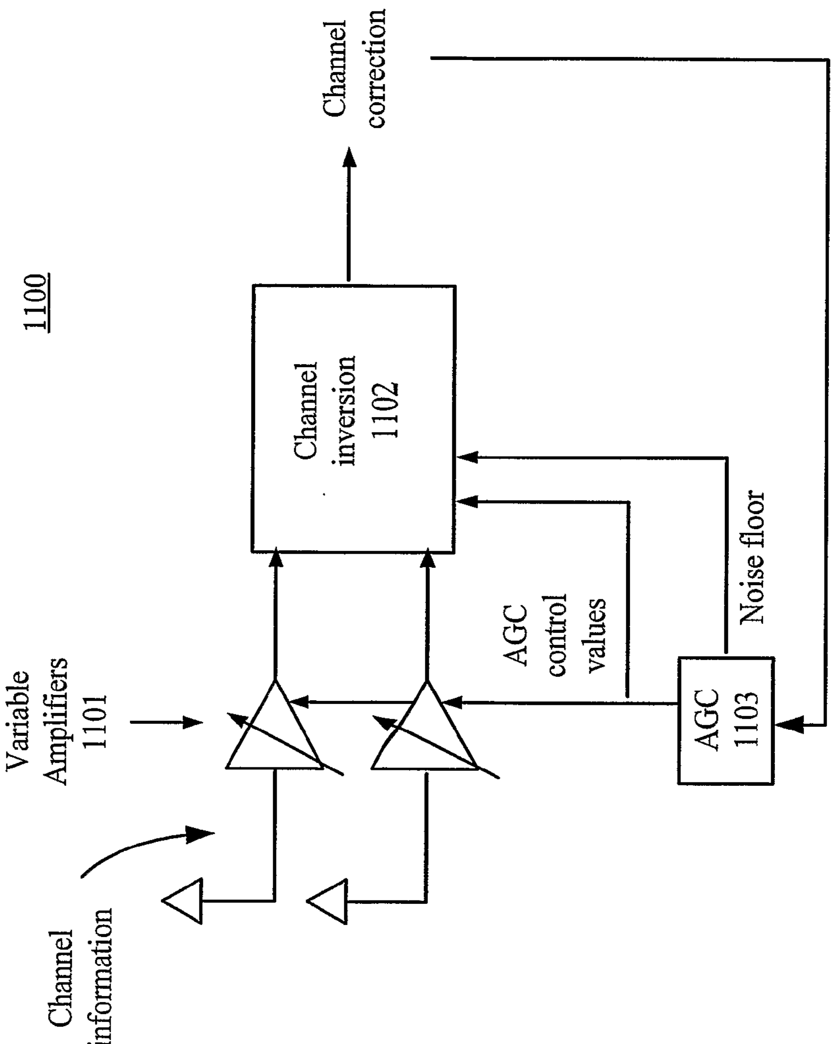
1100
Channel correction
Channel inversion 1102
Variable Amplifiers 1101
AGC control values
Noise floor
AGC 1103
Channel information

Figure 11

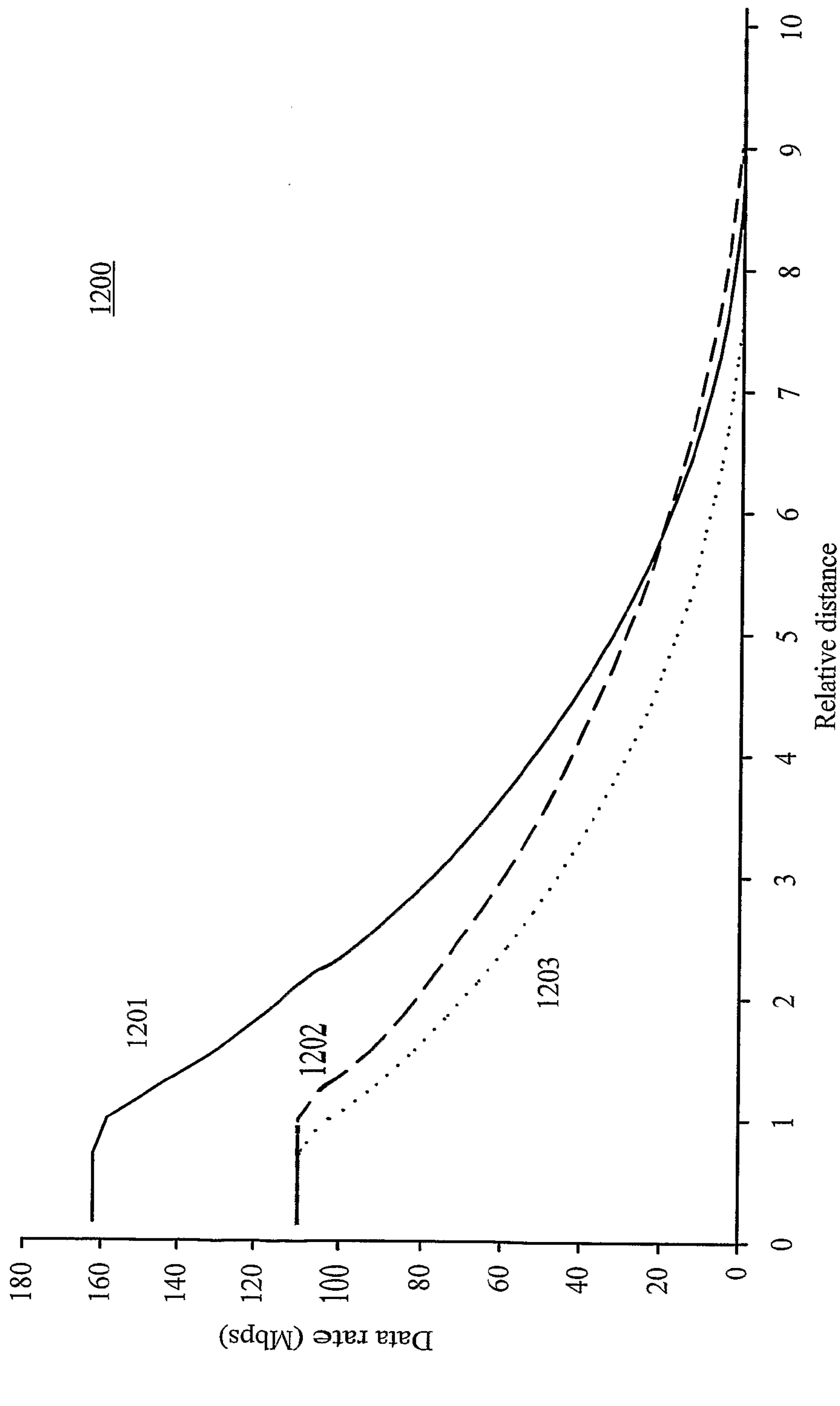
1200
1201
1202
1203
Data rate (Mbps)
180
160
140
120
100
80
60
40
20
0
Relative distance
0
1
2
3
4
5
6
7
8
9
10

Figure 12

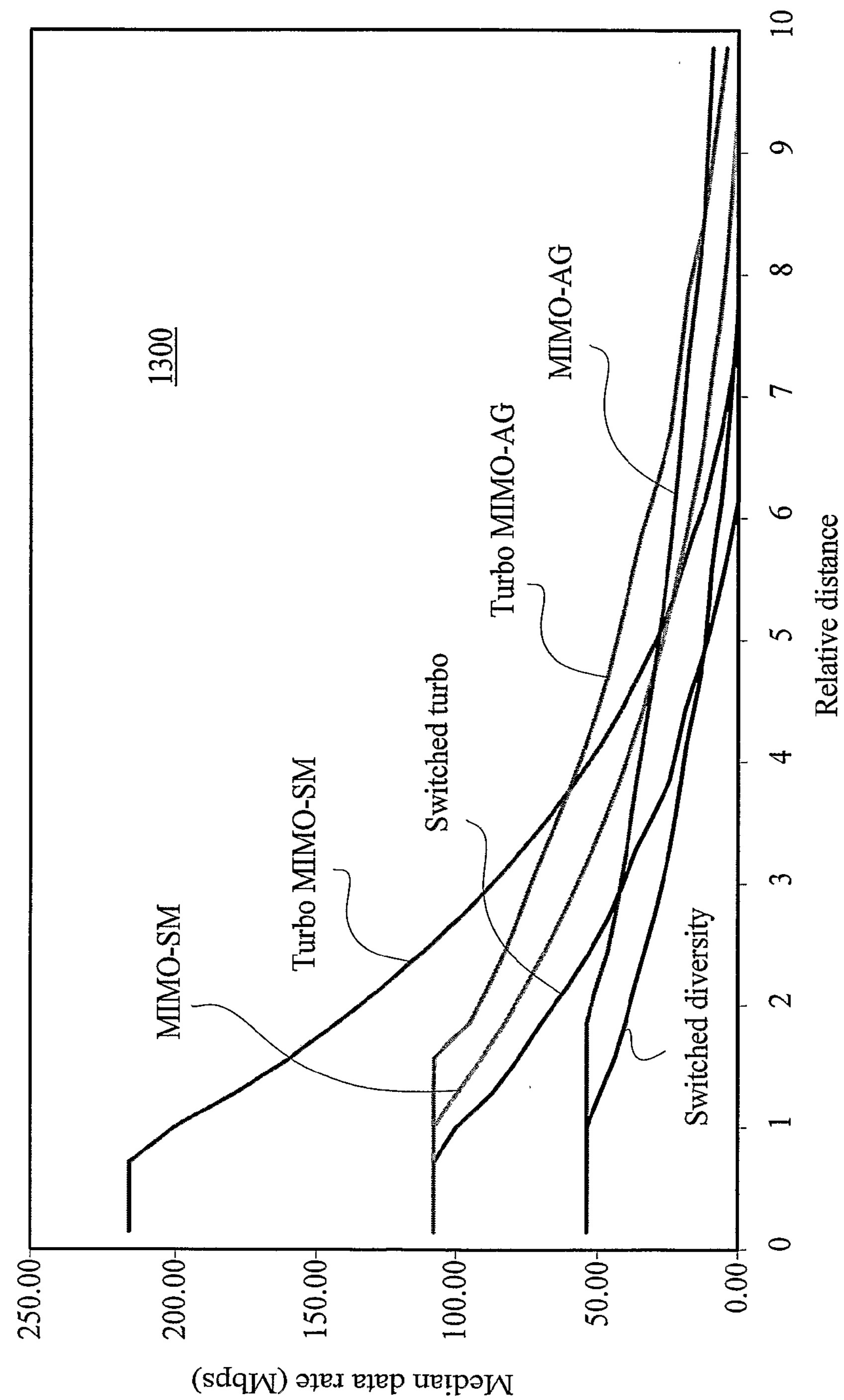
1300
MIMO-SM
Turbo MIMO-SM
Switched turbo
Turbo MIMO-AG
MIMO-AG
Switched diversity
Median data rate (Mbps)
250.00
200.00
150.00
100.00
50.00
0.00
0
1
2
3
4
5
6
7
8
9
10
Relative distance

Figure 13

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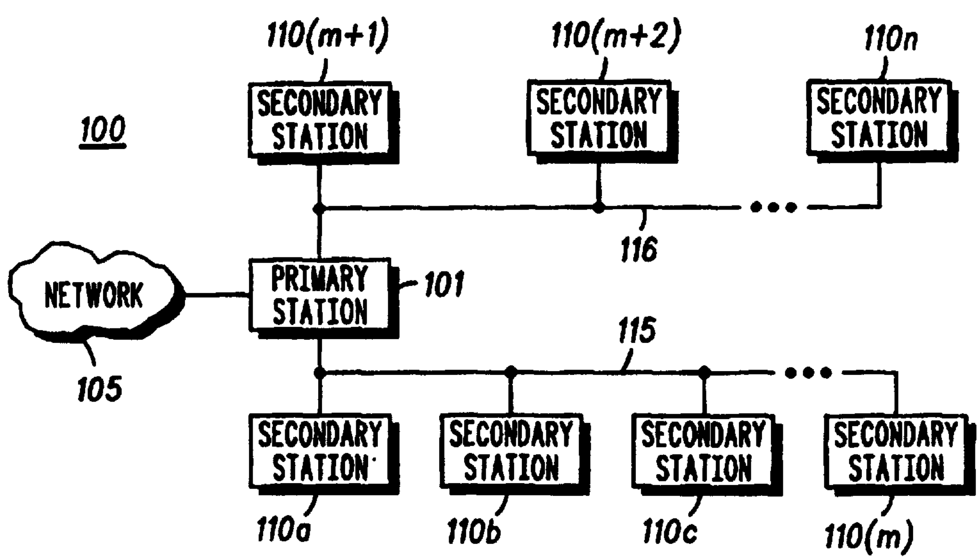

(57) Abstract

An apparatus (101, 110) and method for adaptive forward error correction in a data communication system (100) provides for dynamically changing forward error correction parameters based upon communication channel conditions. Data having a current degree of forward error correction is received (305), and a channel parameter is monitored (310). A threshold level for the channel parameter is determined (315), and the monitored channel parameter is compared to the threshold level (320). When the channel parameter is not within a predetermined or adaptive variance of the threshold level, a revised forward error correction parameter having a greater or lesser degree of forward error correction capability is selected (330, 340, 350, 360), and the revised forward error correction parameter is transmitted (370). The device receiving the revised forward error correction parameter, such as a secondary station (110), then transmits data encoded utilizing the revised error correction parameter (425).

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DK	Denmark	**LR**	Liberia	**SG**	Singapore		
EE	Estonia						

APPARATUS AND METHOD FOR ADAPTIVE FORWARD ERROR CORRECTION IN DATA COMMUNICATIONS

Field of the Invention

This invention relates, in general, to data communications and data communications systems and devices and, more specifically, to an apparatus and method for adaptive forward error correction in data communications.

Background of the Invention

With the advent of multimedia communications, data transmission has become increasingly complex. For example, multimedia communications applications such as real time transmission of digitally encoded video, voice, and other forms of data, may require new forms and systems for data communication and data transmission. One such new communication system is the CableComm™ System currently being developed by Motorola, Inc. In the CableComm™ System, a hybrid optical fiber and coaxial cable is utilized to provide substantial bandwidth over existing cable lines to secondary stations such as individual, subscriber access units, for example, households having new or preexisting cable television capability. These coaxial cables are further connected to fiber optical cables to a central location having centralized, primary (or "head end") controllers or stations having receiving and transmitting capability. Such primary equipment may be connected to any variety of networks or other information sources, from the Internet, various on line services, telephone networks, to video/movie subscriber service. With the CableComm™ System, digital data may be transmitted both in the downstream direction, from the primary station or controller (connected to a network) to the secondary station of an individual user (subscriber access unit), and in the upstream

direction, from the secondary station to the primary station (and to a network).

In the CableComm™ System, downstream data is currently intended to be transmitted using 64 quadrature amplitude modulation ("QAM") at a rate of 30 M bps (megabits per second), over channels having 6 MHz bandwidth in the frequency spectrum of 50 - 750 MHz. Anticipating asymmetrical requirements with large amounts of data tending to be transmitted in the downstream direction rather than the upstream direction, less capacity is provided for upstream data transmission, using $\pi/4$ differential quadrature phase shift keying ($\pi/4$-DQPSK) modulation in the frequency band from 5 - 42 MHz with a symbol rate of 384 k symbols/sec with 2 bits/symbol. In addition, the communication system is designed to have a multipoint configuration, i.e., many end user secondary stations (also referred to as subscriber access units) transmitting upsteam to a primary station, with one or more primary stations transmitting downstream to the secondary stations. The communication system is also designed for asynchronous transmission, with users transmitting and receiving packets of encoded data, such as video or text files. In addition, it is also highly likely that transmission may be bursty, with various users receiving or transmitting data at indeterminate intervals over selected channels in response to polling, contention, or other protocols from the primary station, rather than transmitting a more continuous and synchronous stream of information over a dedicated or circuit switched connection.

For such asynchronous data transmission, it is highly desirable to organize data into recognizable formats or packets for reliable detection by the receivers of the primary station or the secondary station. In the CableComm™ System, the initial portion (or preamble) of the data packet contains timing or synchronization information for accurate data transmission. Following the timing information is encoded

data, which may be encoded for both security (encryption) and for error correction. Following the encoded data are error correction information (as encoded bits) and also additional error detection information in the form of cyclic redundancy check (CRC) bits. One difficulty with inclusion of such error correction information is that such inclusion increases the overall packet size, adding overhead for data transmission and correspondingly decreasing data throughput. Secondly, the inclusion of such error correction information typically increases the system response time or latency, due to the time which may be consumed in the error correction encoding and decoding processes. In addition, there may be situations, such as low noise conditions, in which inclusion of such error correction information may be unnecessary, and higher data throughput may be achieved without the additional overhead of error correction information. Various prior art methods for providing error correction capability, however, typically provided only for a fixed error correction capability, without regard for other opportunities to increase data throughput, for low noise conditions, or for needs to decrease response latency. Accordingly, a need has remained for an apparatus and method to provide for adaptive and flexible error correction capability, providing sufficient error correction for accurate data reception while simultaneously providing for overhead minimization for increased data throughput, and for such an apparatus and method to respond and adapt to potentially changing and variable communication channel conditions.

Brief Description of the Drawings

FIG. 1 is a block diagram illustrating a communication system in accordance with the present invention.

FIG. 2 is a block diagram illustrating a primary station apparatus in accordance with the present invention.

FIG. 3 is a block diagram illustrating a secondary station apparatus in accordance with the present invention.

FIG. 4 is a flow chart illustrating channel monitoring and forward error correction parameter adjustment in accordance with the present invention.

FIG. 5 is a flow chart illustrating forward error correction adjustment and data transmission in accordance with the present invention.

Detailed Description of the Invention

As mentioned above, a need has remained for an apparatus and method to provide for adaptive and flexible error correction capability. The apparatus and method in accordance with the present invention provides such adaptive and flexible error correction capability, providing sufficient error correction for accurate data reception, and also providing for overhead minimization for increased data throughput. The apparatus and method of the present invention is also able to respond and adapt to potentially changing and variable communications channel conditions, such as changes in noise conditions and error rates.

FIG. 1 is a block diagram illustrating a communication system 100 in accordance with the present invention. As illustrated in FIG. 1, a primary station 101, also referred to as a primary transceiver or a primary device, is coupled to a plurality of secondary stations 110_a through 110_n, via communications (or communication) media 115 and 116. In the preferred embodiment, communications media 115 and 116 are hybrid optical fiber and coaxial cable. In other embodiments, the communications media, such as communications media 115 and 116, may be coaxial cable, fiber optic cable, twisted pair wires, and so on, and may also include air, atmosphere or space for wireless and satellite communication. The primary station 101 is also coupled to a network 105, which may include

networks such as the Internet, on line services, telephone and cable networks, and other communication systems. The secondary stations 110_a through 110_n are illustrated in FIG. 1 as connected to the primary station 101 on two segments or branches of a communications medium, such as communications media 115 and 116. Equivalently, the secondary stations 110_a through 110_n may be connected to more than one primary station, and may be connected to a primary station (such as primary station 101) utilizing more or fewer branches, segments or sections of any communications medium.

Continuing to refer to FIG. 1, in the preferred embodiment, the communications medium, such as communications media 115 and 116, has or supports a plurality of communications channels. For ease of reference, the communications channels in which a primary station, such as the primary station 101, transmits information, signals, or other data to a secondary station, such as secondary station 110_n, are referred to as downstream channels or downstream communication channels. Also for ease of reference, the communications channels in which a secondary station, such as secondary station 110_n, transmits information, signals, or other data to a primary station, such as primary station 101, are referred to as upstream channels or upstream communication channels. These various upstream and downstream channels may, of course, be the same physical channel or may be separate physical channels, for example, through time division multiplexing or frequency division multiplexing. These various channels may also be logically divided in other ways, in addition to upstream and downstream directions. As mentioned above, in the preferred embodiment of the CableComm™ System, the communications medium is hybrid fiber coaxial cable, with downstream channels in the frequency spectrum (or band) of 50 - 750 MHz, and with upstream channels in the frequency spectrum of 5 - 42 MHz.

FIG. 2 is a block diagram illustrating a primary station 101 in accordance with the present invention. The primary station 101 is coupled to a communication medium 114 for upstream and downstream communication to one or more secondary stations (not illustrated), and is coupleable to a network, such as the Internet, through a network interface 119. The primary station includes a processor arrangement 120 which is connected to a plurality of channel interfaces, channel interface 125_a through channel interface 125_n, for communication over the communication medium 114. The processor arrangement 120 includes a master controller 121 having or connected to memory 122, and one or more additional processors 130_{a1} through 130_{n2} and corresponding associated memories 131_{a1} through 131_{n2}. In the preferred embodiment, the master controller 121 is a Motorola M68040 processor, and the memory 122 is 16 MB RAM. The master controller 121 performs a variety of higher level functions in the preferred embodiment, such as spectrum management, routing functions, management of secondary stations, and communication protocol management (such as SNMP management). The master controller 121 is connected to a plurality of other processors, collectively referred to as processors 130 and separately illustrated as processor 130_{a1}, processor 130_{a2}, through processor 130_{n1} and processor 130_{n2}. Each of these processors, processor 130_{a1}, processor 130_{a2}, through processor 130_{n1} and processor 130_{n2}, is also coupled to or contains corresponding memory circuits, memory 131_{a1}, memory 131_{a2}, through memory 131_{n1} and memory 131_{n2}. In the preferred embodiment, each of these processors 130 are also Motorola M68040 processors, while the corresponding memory circuits, memory 131_{a1} through memory 131_{n2}, are 4 MB RAM. In the preferred embodiment, the processors 130 perform such functions related to upstream and downstream data protocols, such as sending a poll message or an acknowledgment message downstream. Each of these

processors 130_{a1} through 130_{n2} of the processor arrangement 120 are connected to corresponding receivers and transmitters of the channel interfaces, channel interface 125_a through channel interface 125_n (collectively referred to as channel interfaces 125), namely, receiver 135_a through receiver 135_n (collectively referred to as receivers 135) and transmitter 136_a through transmitter 136_n (collectively referred to as transmitters 136). In the preferred embodiment, depending upon the functions implemented, each of the receivers 135_a through 135_n may include a Motorola M68302 processor, a Motorola 56000 series digital signal processor, a ZIF SYN integrated circuit, and an LSI Logic L64714 (Reed-Solomon decoder), for demodulation and for decoding forward error correction and cyclic redundancy checks. In the preferred embodiment, also depending upon the functions implemented, each of the transmitters 136_a through 136_n may include a Motorola M68302 processor, a Motorola 56000 series digital signal processor, a ZIF SYN integrated circuit, and an LSI Logic L64711 (Reed-Solomon encoder), for modulation and for coding for forward error correction and cyclic redundancy checks. As a consequence, as used herein, the channel interfaces 125 may be considered to perform the functions of data and other signal reception and transmission, regardless of the specific hardware implementations and additional functions which may or may not be implemented. The various memories illustrated, such as memory 122 or 131_{a1}, may also be embodied or contained within their corresponding processors, such as master controller 121 or processor 130_{a1}. The functions of these various components with respect to the present invention are explained in greater detail below with reference to FIGs. 4 and 5.

FIG. 3 is a block diagram illustrating a representative secondary station 110_n (of the plurality of secondary stations 110) in accordance with the present invention. The secondary station 110_n includes a processor (or processor arrangement)

150, with the processor 150 having or coupled to a memory 155. In the preferred embodiment, the processor 150 is a Motorola M68302 processor (also known as an integrated multiprotocol processor), and the memory 155 is 256 K RAM. The processor 150 is coupled to an interface 170, such as an ethernet port or an RS232 interface, for connection to a computer, a workstation, or other data terminal equipment ("DTE"). The processor 150 is also coupled to a channel interface 160 for communication over the communication medium 114. The channel interface 160, in the preferred embodiment, depending upon the functions implemented, includes a Motorola M68HC11 integrated circuit, a ZIF SYN integrated circuit, a Broadcom BCM3100 QAMLink integrated circuit, a Motorola TxMod integrated circuit, and LSI Logic L64711 and L64714 integrated circuits, and performs such functions as forward error correction encoding and decoding, QAM demodulation (for downstream reception), QPSK modulation (for upstream transmission), transmit level and frequency adjustment, for data and other signal reception and transmission. As a consequence, as used herein, the channel interface 160 may be considered to perform the functions of data and other signal reception and transmission, regardless of the specific hardware implementations and additional functions which may or may not be implemented. The memory illustrated as memory 155 may also be embodied or contained within the corresponding processor 150. The additional functions of these components of a secondary station 110_n with respect to the invention are also described in greater detail below with reference to FIGs. 4 and 5.

As mentioned above, the upstream channels of the communication medium, in the preferred CableComm™ System, are in a frequency range between 5 and 42 MHz and may be susceptible to interference from typical noise sources. Forward error correction is preferably employed on the upstream channels as a way of compensating for data

transmission errors which may have been caused by noise or other distortions. Forward error correction comprises an error correcting code that is added to the user data to allow a receiver to correct certain types and sizes of errors that may have occurred during the transmission of the data. The transmitting unit, such as the processor 150 and channel interface 160 of a secondary station 110_n, generates the error correcting code from the user data, and appends the encoded data onto the user data during transmission. The receiving unit, such as receiver 135_n and processor 130_{n2} of the primary station 101, uses the encoded data to detect received errors and to correct detected errors. As a consequence, the receiving unit should know, prior to the receipt of actual data, what type of error correcting code is to be employed by the transmitting unit, for proper decoding and error correction. This may be typically done by prior agreement (e.g., during initial set up or configuration of the communication system), or during a negotiation "handshake" during establishment of the communications link.

In addition, there are many types of error correcting codes, which are generally categorized as either convolutional codes, which correct random bit errors, and block codes, which correct burst errors. Two or more error correcting codes may be used together to obtain a total error correcting capability or power that is greater than the sum of the capabilities of the individual codes, and are typically referred to as "concatenated" codes. A popular concatenated code uses a convolutional "inner" code and a block "outer" code. The performance of a block error correcting code, moreover, may often be increased using an "interleaving" technique, in which data which may be subject to a burst error is spread out over multiple blocks, thereby providing each block code a higher probability of correcting a small portion of the large burst error. The correcting power of an interleaving technique is

determined or measured as a function of interleaver depth. Trellis coding techniques may also be utilized.

The preferred embodiment utilizes a Reed-Solomon error correcting code for forward error correction on the upstream channels, without additional convolutional coding and interleaving. The Reed-Solomon error correcting code is known and is a block code, such that the error correcting code is computed over a block of data having a fixed size. A Reed-Solomon code is typically specified by a parameter pair (n, k), in which "n" is the code word size and "k" is the block size, such that an n-byte code word consists of k data bytes and (n - k) redundancy bytes (which represent the error correcting code information). The maximum number of symbol errors that can be corrected by a Reed-Solomon code is $t = (n - k)/2$, where a symbol is typically one 8-bit byte. A commonly used Reed-Solomon code is a (128,122) code, where the code word size is 128 bytes, each code word consists of 122 data bytes and 6 redundancy bytes, enabling a decoder to correct up to three byte errors in each 128 byte code word. In addition to a Reed-Solomon code, other error correcting codes and encryption algorithms may also be used.

In a typical prior art forward error correction implementation, the forward error correction parameters are set to a predetermined and fixed value to compensate for a particular level of noise on the communications channel. If the noise level on the communications channel becomes excessive, such that the noise exceeds the ability of the forward error correction to correct transmission errors, the data will be received in error. In that case, the data must be retransmitted or, in a worst case situation, the communications channel may no longer be usable. In either situation, data throughput is significantly decreased (or eliminated).

Forward error correction parameters, however, illustrate a balance between the amount of overhead added by the error correcting code itself (which utilizes space which could have

been used for data and therefore decreases data throughput), on the one hand, and the amount of error correction needed due to channel conditions (which may serve to increase data throughput through avoidance of retransmission), on the other hand. In the preferred embodiment, to maximize throughput of user data over a given communications channel, the optimal error correcting methodology would utilize precisely enough error correction to compensate for the existing noise level, and no more or less. Any more error correcting capability lowers throughput due to excessive overhead from transmitting the error correcting code information, while any less lowers the throughput due to the overhead caused by retransmission of data received in error. The level of noise on a communications channel, however, may vary over time, rendering a selection of a fixed set of forward error correction parameters less than optimal at any given time. One prior art method, as mentioned above, selects a fixed set of forward error correction parameters to compensate for a typical or anticipated noise level, but ceases to use the communication channel when the noise becomes excessive. This prior art method of utilizing fixed error correction code parameters is unsuitable for situations in which the number of available channels is limited, in which case it would be preferable to maintain a channel at a reduced throughput level rather than eliminate use of the channel altogether.

Secondly, another objective of the communication system of the preferred embodiment concerns minimizing the amount of throughput delay introduced by the communications equipment. Throughput delay in a polled protocol, for example, may be defined as the amount of time between the sending of a poll message prior to forward error correction encoding and the receipt of a response to the poll message following forward error correction decoding. Forward error correcting codes typically introduce such throughput delay due to the processing and computational requirements of error correction

encoding and decoding, and the amount of throughput delay is typically proportional to the error correcting power of the code. For example, the delay introduced by the interleaving/de-interleaving process is proportional to the interleaver depth, and the delay introduced by the Reed-Solomon encoding/decoding process is proportional to the code word size and number of redundancy bytes.

As discussed in greater detail below, the apparatus and method of the present invention provides a means for signaling and changing the forward error correction parameters (typically used on an upstream channel), based upon variable channel quality, such as variable noise or error levels. As a consequence, the preferred embodiment of the present invention provides a mechanism to optimize data throughput for varying noise levels (and corresponding error rates), while simultaneously providing a mechanism to decrease throughput delay as needed or as desired.

FIG. 4 is a flow chart illustrating channel monitoring and forward error correction parameter adjustment in accordance with the present invention. Beginning with start step 300, the method proceeds to receive encoded data having a first or initial degree of forward error correction, from a plurality of degrees of forward error correction, in step 305. The plurality of degrees of forward error correction result from the variable levels of correcting capability associated with various codes and with various parameters of such codes. For example, different error correcting capabilities result from the specification of different error correcting parameters such as (n, k) parameters, e.g., (128, 122), (200, 196), or (128, 124), from inclusion of different types of error correcting codes, such as concatenation of codes or inclusion of interleaving (with a specified depth), and by specification of any parameters associated with such codes. In the preferred embodiment, the initial degree of forward error correction may be established during an initial registration process, when a

secondary station establishes a communication link with a primary station. Typically in the preferred embodiment, the primary station polls individual secondary stations, where each poll message contains a secondary station identifier, an upstream channel number, and the parameters to be used for forward error correction on an upstream response. As explained with reference to FIG. 5, when the secondary station receives the poll on the downstream channel, it responds on the upstream channel designated in the poll message using the forward error correction parameters also specified in the poll message.

Continuing to refer to FIG. 4, following receipt of encoded data having an initial degree of forward error correction in step 305, channel parameters are monitored in step 310, such as monitoring packet error rates, bit error rates, noise levels (such as levels of impulse noise or ingress noise), other interference, or other parameters or factors which could be correlated with channel quality, error rates, and a desired or necessary degree of error correction capability. For example, monitoring an error rate may comprise monitoring a set of error rate parameters of a plurality of sets of error rate parameters in which the plurality of sets of error rate parameters consist of any of a plurality of combinations of a bit error rate, a packet error rate, a burst error rate, a block error rate, and a frame error rate. Next, a threshold level is determined in step 315, such as a threshold level of packet errors or bit errors. This threshold level may be predetermined, may be set at a default value, or may be adaptive and take on various values depending upon the allowable amount of transmission error. For example, under conditions in which few errors will be allowed, the threshold level may be comparatively low. In circumstances in which throughput latency may be more significant and more errors may be allowable, the threshold level may be comparatively high. Next, in step 320, the channel parameter (monitored in

step 310) is compared to the threshold level, step 320, such that if the channel parameter is not within a predetermined or adaptive tolerance or variance of the threshold level (i.e., is not equal to the threshold level plus or minus an allowable tolerance (or variance)), then the degree of forward error correction utilized will be revised, as explained below. The tolerance or variance level may be either predetermined, such as a fixed variance, or adaptive, such as a variance which is changeable over time. If the channel parameter is greater than the threshold level (plus the allowable tolerance, if any) in step 330, indicating that the channel (and its associated noise and distortions) has a comparatively high quality and is causing fewer errors than are capable of being corrected by the current error correcting code, then a revised forward error correcting parameter is selected which has a lower degree of forward error correction capability, i.e., is capable of correcting fewer errors, step 340. In such a case, a lower or lesser degree of forward error correction capability is selected to decrease the overhead associated with utilizing greater error correction capability, when such greater error correction capability is unnecessary because the channel parameter indicates fewer errors needing correction. Conversely, if the channel parameter is less than the threshold level (minus the allowable tolerance, if any) in step 350, indicating that the channel (and its associated noise and distortions) has a comparatively low or poor quality and is causing more errors than are capable of being corrected by the current error correcting code, then a revised forward error correcting parameter is selected which has a higher degree of forward error correction capability, i.e., is capable of correcting more errors, step 360. In this case, a greater or higher degree of forward error correction capability is selected to decrease the overhead associated with retransmission of an entire data packet due to excessive errors, when such greater error correction capability is

necessary because the channel parameter indicates more errors needing correction. Eqivalently, depending upon the choice of channel parameter employed, such as noise levels or error rates, the comparative steps 330 and 350 may be reversed or modified. For example, if the selected channel parameter is an error rate, such that if the threshold error rate level with a selected tolerance is exceeded in step 330, then a revised forward error correcting parameter is selected which has a higher degree of forward error correction capability in step 340, with corresponding modifications of steps 350 and 360. Following steps 340 and 360, the revised forward error correction parameter is transmitted, step 370, for example, to a particular secondary station, such as secondary station 110_m. As this process may be both repeated for each connected (or active) secondary station 110_a through 110_n, and repeated over time (as conditions may vary), it is anticipated that different revised forward error correction parameters will be determined and transmitted to the different secondary stations, and over time, to the same secondary station. The various comparative steps 330 and 350 may also include variance or tolerance levels, such that forward error correction might not be revised unless the channel parameter differs from the threshold level by a predetermined amount or variance, to avoid any interruptions or delay due to comparatively minor variances and correspondingly minor changes in forward error correction capability. In addition, as the error parameters may vary over time, it may be desirable to utilize an average value of error parameters, rather than instantaneous values. The process ends (return step 380) following step 370, or when the channel parameter is equal to the threshold level, i.e., the channel parameter is not greater than the threshold level in step 330 and the channel parameter is not less than the threshold level in step 350, or is otherwise within the allowable tolerance or variance from the threshold level.

In the preferred apparatus embodiment illustrated in FIG. 2, the method illustrated in FIG. 4 may be programmed and stored, as a set of program instructions for subsequent execution, in the primary station 101 and, more particularly, in each of the processors 130 (with their associated memories 131), utilizing packet or bit error data from the forward error correction decoding performed by the corresponding receivers 135. To the extent that adjustable and dynamic forward error correction capability is necessary or desirable in the downstream direction, the method illustrated in FIG. 4 also may be programmed and stored, also as a set of program instructions for subsequent execution, in the processor 150 and memory 155 of a secondary station 110, such as secondary station 110_n illustrated in FIG. 3.

In summary, FIGs. 1 and 4 illustrate a method for adaptive forward error correction in a data communication system 100, the data communication system having a communications medium (such as 114, 115 or 116), with the communications medium having a plurality of communications channels. The method then comprises: (a) receiving encoded data over a first communications channel of the plurality of communications channels, the encoded data having a first degree of forward error correction of a plurality of degrees of forward error correction (step 305); (b) monitoring a channel parameter (such as error rate, ingress noise or impulse noise) of the first communications channel to form a monitored parameter (step 310); (c) determining a threshold level of the channel parameter (step 315); (d) comparing the monitored parameter with the threshold level (step 320); (e) when the monitored parameter is not within a variance (either predetermined or adaptive) of the threshold level, changing the first degree of forward error correction to a second degree of forward error correction of the plurality of degrees of forward error correction (steps 330, 340, 350 and 360); and (f) transmitting a forward error correction revision parameter on

a second communications channel of the plurality of communications channels, the forward error correction revision parameter corresponding to the second degree of forward error correction (step 370). The various first and second communications channels, of course, may be the same or different logical or physical channels (such as time or frequency division multiplexed channels).

FIG. 5 is a flow chart illustrating forward error correction adjustment and data transmission in accordance with the present invention. In the preferred embodiment, this forward error correction adjustment and data transmission methodology, for upstream data transmission, would also be implemented as a set of program instructions stored in the processor 150 and memory 155 of a secondary station 110. Correspondingly, to the extent that adjustable and dynamic forward error correction capability is necessary or desirable in the downstream direction, the method illustrated in FIG. 5 also may be programmed and stored, also as a set of program instructions for subsequent execution, in a primary station 101, and more particularly, in its processor arrangement 120 (with associated memories), utilizing the forward error correction encoding capability of the associated transmitters 136.

Referring to FIG. 5, beginning with start step 400, data having an initial degree of forward error correction (of a plurality of degrees of forward error correction) is transmitted on a communication channel, step 405. The initial degree of forward error correction may be either predetermined in a secondary station 110_n, such as currently fixed in its software, or may be signaled or otherwise transmitted from a primary station 101, such as in a poll message as mentioned above. The communication channel is then monitored for reception of a revised forward error correction parameter, step 410, which may, for example, be contained in a specific poll message to a secondary station or

may be broadcast in a message to all secondary stations. In addition, while the preferred embodiment utilizes signaling in a link layer through specific poll or broadcast messages, such signaling may be performed at any layer in a communication protocol, including the physical layer, the network layer, or in a software download (from a primary station 101 to a secondary station 110_n). If a revised degree of forward error correction is not received in step 415, for example, because a received poll continues to indicate the current degree of forward error correction or because a specific poll or other message containing a revised parameter was not received, then the method continues to transmit data at the then current degree (such as the initial degree) of forward error correction, step 420, and continues to monitor the communication channel for a revised parameter, returning to step 410. If a revised degree of forward error correction is received in step 415, then data will be transmitted utilizing the revised degree of forward error correction indicated by the revised forward error correction parameter, step 425, and the process ends, return step 430. As mentioned above with regard to FIG. 4, this process may be repeated for each secondary device and over time (as conditions may vary), and it is also anticipated that different revised forward error correction parameters will be received by the different secondary stations, and over time, by the same secondary station.

In summary, FIG. 5 illustrates a method for adaptive forward error correction in a data communications system 100, the data communications system 100 having a communications medium (114, 115 or 116), with the communications medium having a plurality of communications channels. The method then comprises: (a) transmitting encoded data over a first communications channel of the plurality of communications channels to form transmitted encoded data, the transmitted encoded data having an initial degree of forward error correction of a plurality of degrees of forward

error correction (step 405); (b) monitoring a second communications channel of the plurality of communications channels for a forward error correction revision parameter (step 410); (c) determining whether the forward error correction revision parameter indicates a revised degree of forward error correction of the plurality of degrees of forward error correction (step 415); (d) transmitting encoded data having the initial degree of forward error correction when the forward error correction revision parameter does not indicate the revised degree of forward error correction (step 420); and (e) transmitting encoded data having the revised degree of forward error correction when the forward error correction revision parameter indicates the revised degree of forward error correction (step 425). Also, as noted above, the various first and second communications channels may be the same or different logical or physical channels.

Also in summary, FIGs. 2 and 4 illustrate an apparatus 101 for adaptive forward error correction in a data communications system 100, the data communications system 100 having a communications medium (114, 115 or 116), with the communications medium having a plurality of communications channels. The apparatus 101 then comprises, first, a channel interface 125 coupleable to the communications medium 114 for transmission of encoded data on a first communications channel of the plurality of communications channels to form transmitted encoded data and for reception of encoded data on a second communications channel of the plurality of communications channels to form received encoded data, with the received encoded data having a first degree of forward error correction of a plurality of degrees of forward error correction; and second, a processor arrangement 120 coupled to the channel interface 125, the processor arrangement 120 responsive through a set of program instructions to monitor a channel parameter of the second communications channel to form a monitored

parameter; the processor arrangement further responsive to compare the monitored parameter with a threshold level of the channel parameter; and when the monitored parameter is not within a variance (or tolerance level) of the threshold level, the processor arrangement is further responsive to change the first degree of forward error correction to a second degree of forward error correction of the plurality of degrees of forward error correction, and to transmit via the channel interface a forward error correction revision parameter on the first communications channel, the forward error correction revision parameter corresponding to the second degree of forward error correction. As illustrated above, the apparatus may be embodied within a primary station 101, which also may be referred to as a primary device or primary transceiver. The channel interface, such as channel interface 125_n, may also further comprise a receiver and a transmitter, such as receiver 135_n and a transmitter 136_n. The processor arrangement 120 may also further comprise: a first processor coupled to the channel interface 125, such as processor 130_{n1}; a second processor coupled to the channel interface 125, such as processor 130_{n2}; and a master controller coupled to the first processor and to the second processor, such as master controller 121.

Also in summary, FIGs. 3 and 5 illustrate an apparatus for adaptive forward error correction in a data communications system 100, the data communications system 100 having a communications medium, with the communications medium having a plurality of communications channels. The apparatus comprises, first, a channel interface 160 coupleable to the communications medium 114 for transmission of encoded data on a first communications channel of the plurality of communications channels to form transmitted encoded data and for reception of encoded data on a second communications channel of the plurality of communications channels to form received encoded data; a

processor (or processor arrangement) 150 (including processor 150 with memory 155) coupled to the channel interface 160, the processor arrangement 150 responsive through a set of program instructions to set the transmitted encoded data to a current (or initial) degree of forward error correction of a plurality of degrees of forward error correction, the processor arrangement further responsive to monitor the second communications channel for reception of a forward error correction revision parameter and to determine whether the forward error correction revision parameter indicates a revised degree of forward error correction of the plurality of degrees of forward error correction; the processor arrangement further responsive to transmit encoded data having the current (or initial) degree of forward error correction when the forward error correction revision parameter does not indicate the revised degree of forward error correction and to transmit encoded data having the revised degree of forward error correction when the forward error correction revision parameter indicates the revised degree of forward error correction. As mentioned above, the processor arrangement may be processor arrangement 120, processor 150 or may also be the processor 150 coupled to the memory 155.

The forward error correction parameters carried in the downstream poll in the preferred embodiment may specify the type or types of error correcting codes and the parameters for each error correcting code. The parameters for forward error correction may be specified in a variety of ways, depending upon the chosen embodiment and the types of codes to be utilized. Utilizing a Reed-Solomon code, the (n, k) parameters may be directly specified, for example, using two bytes such as (128, 122). In the preferred embodiment, to decrease the overhead content in poll and other messages, a one byte parameter is utilized as an index to a table (look up table) containing up to 256 variations or combinations of Reed-

Solomon (n, k) parameters. The 256 combinations of (n, k) parameters are selected on the basis of which are most likely to be utilized in the selected communications system, such as the CableComm™ System. For example, rather than transmitting two bytes of information specifying (128, 122), in the preferred embodiment, a parameter consisting of one byte of information is transmitted as an index (or pointer) to a look up table stored in memory, which is then translated or converted to the selected (n, k) combination, such as (128, 122). Also in the preferred embodiment, forward error correction may be disabled altogether if a channel has sufficiently low noise, and each type of error correction (e.g., convolutional coding, block coding, concatenation, and interleaving) can be individually enabled or disabled utilizing a specified set of operating parameters.

While carrying the forward error correction information in each downstream poll is the preferred method for the CableComm™ System, an alternative method is to use a special downstream message that is transmitted only when error correcting power is revised. This eliminates the overhead of carrying this information in the downstream polls, which may be frequently sent messages. Moreover, in addition to specifying forward error correction parameters, the primary station could also specify, both initially and as may be subsequently revised, analog parameters to be used for the upstream transmission, such as a modulation mode, carrier frequency, bit rate, baud rate, and bandwidth for each carrier. Improved throughput may be realized by changing the analog parameters, such as modulation mode, instead of the forward error correction, where the amount of overhead that would be added by the forward error correction is greater than the throughput loss caused by a slower but more robust modulation mode. The primary station would then be able to vary the analog parameters, forward error correction, or both, in order to compensate for channel impairments.

Although the preferred embodiment utilizes a poll/response protocol, the invention may also be applied to non-polled protocols, and may use the same or different signaling techniques for dynamically adapting the types of forward error correction and the forward error correction parameters. The invention could also be used where dial-up or other lines are used as the upstream channel. Even though the dial-up line is a dedicated channel, channel characteristics over a public switched network may vary over time and also may vary from connection to connection. Thus, the dynamic adaptive forward error correction could be used to improve throughput over such a dedicated channel.

The ability to dynamically adapt the level of forward error correction for individual channels, in accordance with the present invention, provides several significant advantages. First, it allows continued use of channels that otherwise would have been vacated of all traffic in an implementation which used the prior art fixed level of forward error correction. Even though the throughput rate for that given channel is diminished due to the increased overhead of a more powerful error correcting code, the overall throughput of the communication system is increased through the utilization of an otherwise unacceptable channel. Second, the apparatus and method of the present invention allows the level of error correction to be tailored for each channel, so that a "clean" channel does not carry a greater amount of overhead or introduce a greater amount of throughput delay than that required to compensate for the specific and actual level of noise, rather than a predetermined or anticipated level of noise.

From the foregoing, it will be observed that numerous variations and modifications may be effected without departing from the spirit and scope of the novel concept of the invention. It is to be understood that no limitation with respect to the specific methods and apparatus illustrated

herein is intended or should be inferred. It is, of course, intended to cover by the appended claims all such modifications as fall within the scope of the claims. The invention is further defined by the following claims.

We claim:

1. A method for determining forward error correction parameters for adaptive forward error correction in a data communication system, the data communication system having a communications medium, the communications medium having a plurality of communications channels, the method comprising:

(a) receiving encoded data over a first communications channel of the plurality of communications channels, the encoded data having a first degree of forward error correction of a plurality of degrees of forward error correction;

(b) monitoring a channel parameter of the first communications channel to form a monitored parameter;

(c) determining a threshold level of the channel parameter;

(d) comparing the monitored parameter with the threshold level;

(e) when the monitored parameter is not within a variance of the threshold level, selecting a second degree of forward error correction of the plurality of degrees of forward error correction; and

(f) transmitting a forward error correction revision parameter on a second communications channel of the plurality of communications channels, the forward error correction revision parameter corresponding to the second degree of forward error correction.

2. The method for adaptive forward error correction in a data communications system of claim 1, wherein the plurality of degrees of forward error correction are comprised of any of a plurality of combinations of parameters specifying block codes, convolutional codes, concatenated codes, and interleaving depth.

3. The method for adaptive forward error correction in a data communications system of claim 1, further comprising:

(h) transmitting a revised analog parameter on a second communications channel of the plurality of communications channels, wherein the revised analog parameter is comprised of any of a plurality of combinations of parameters specifying a modulation mode, a carrier frequency, a bit rate, a baud rate, and a carrier bandwidth.

4. The method for adaptive forward error correction in a data communications system of claim 1, wherein step (e) further comprises:

(e1) when the monitored parameter as compared to the threshold level indicates that a lesser degree of forward error correction is appropriate, selecting the second degree of forward error correction having a lesser error correction capacity than the first degree of error correction; and

(e2) when the monitored parameter as compared to the threshold level indicates that a greater degree of forward error correction is appropriate, selecting the second degree of forward error correction having a greater error correction capacity than the first degree of error correction.

5. A method for revising forward error correction parameters for adaptive forward error correction in a data communications system, the data communications system having a communications medium, the communications medium having a plurality of communications channels, the method comprising:

(a) transmitting encoded data over a first communications channel of the plurality of communications channels to form transmitted encoded data, the transmitted encoded data having a current degree of forward error correction of a plurality of degrees of forward error correction;

(b) monitoring a second communications channel of the plurality of communications channels for a forward error correction revision parameter;

(c) determining whether the forward error correction revision parameter indicates a revised degree of forward error correction of the plurality of degrees of forward error correction;

(d) transmitting encoded data having the current degree of forward error correction when the forward error correction revision parameter does not indicate the revised degree of forward error correction; and

(e) transmitting encoded data having the revised degree of forward error correction when the forward error correction revision parameter indicates the revised degree of forward error correction.

6. An apparatus for determining forward error correction parameters for adaptive forward error correction in a data communication system, the data communication system having a communications medium, the communications medium having a plurality of communications channels, the apparatus comprising:

a channel interface coupleable to the communications medium for transmission of encoded data on a first communications channel of the plurality of communications channels to form transmitted encoded data and for reception of encoded data on a second communications channel of the plurality of communications channels to form received encoded data, the received encoded data having a first degree of forward error correction of a plurality of degrees of forward error correction; and

a processor arrangement coupled to the channel interface, the processor arrangement responsive through a set of program instructions to monitor a channel parameter of the second communications channel to form a monitored

parameter; the processor arrangement further responsive to compare the monitored parameter with a threshold level of the channel parameter and, when the monitored parameter is not within a variance of the threshold level, the processor arrangement is further responsive to select a second degree of forward error correction of the plurality of degrees of forward error correction and to transmit via the channel interface a forward error correction revision parameter on the first communications channel, the forward error correction revision parameter corresponding to the second degree of forward error correction.

7. The apparatus of claim 6, wherein the plurality of degrees of forward error correction are comprised of any of a plurality of combinations of parameters specifying block codes, convolutional codes, concatenated codes, and interleaving depth.

8. The apparatus of claim 6, wherein the processor arrangement is further responsive to transmit a revised analog parameter on the first communications channel of the plurality of communications channels, wherein the revised analog parameter is comprised of any of a plurality of combinations of parameters specifying a modulation mode, a carrier frequency, a bit rate, a baud rate, and a carrier bandwidth.

9. The apparatus of claim 6, wherein when the monitored parameter as compared to the threshold level indicates that a lesser degree of forward error correction is appropriate, the processor arrangement is further responsive to select the second degree of forward error correction having a lesser error correction capacity than the first degree of error correction; and when the monitored parameter as compared to the threshold level indicates that a greater degree of forward

error correction is appropriate, the processor arrangement is further responsive to select the second degree of forward error correction having a greater error correction capacity than the first degree of error correction.

10. An apparatus for revising forward error correction parameters for adaptive forward error correction in a data communications system, the data communications system having a communications medium, the communications medium having a plurality of communications channels, the apparatus comprising:

a channel interface coupleable to the communications medium for transmission of encoded data on a first communications channel of the plurality of communications channels to form transmitted encoded data and for reception of encoded data on a second communications channel of the plurality of communications channels to form received encoded data; and

a processor arrangement coupled to the channel interface, the processor arrangement responsive through a set of program instructions to set the transmitted encoded data to a current degree of forward error correction of a plurality of degrees of forward error correction, the processor arrangement further responsive to monitor the second communications channel for reception of a forward error correction revision parameter and to determine whether the forward error correction revision parameter indicates a revised degree of forward error correction of the plurality of degrees of forward error correction; the processor arrangement further responsive to transmit encoded data having the current degree of forward error correction when the forward error correction revision parameter does not indicate the revised degree of forward error correction and to transmit encoded data having the revised degree of forward error correction when the forward error correction revision

parameter indicates the revised degree of forward error correction.

FIG. 1

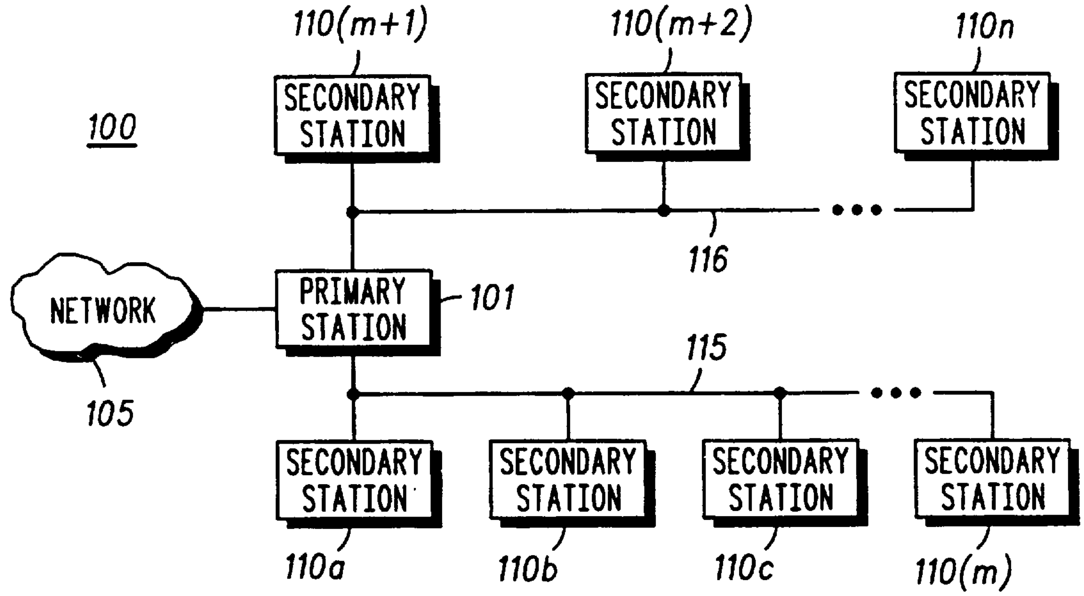
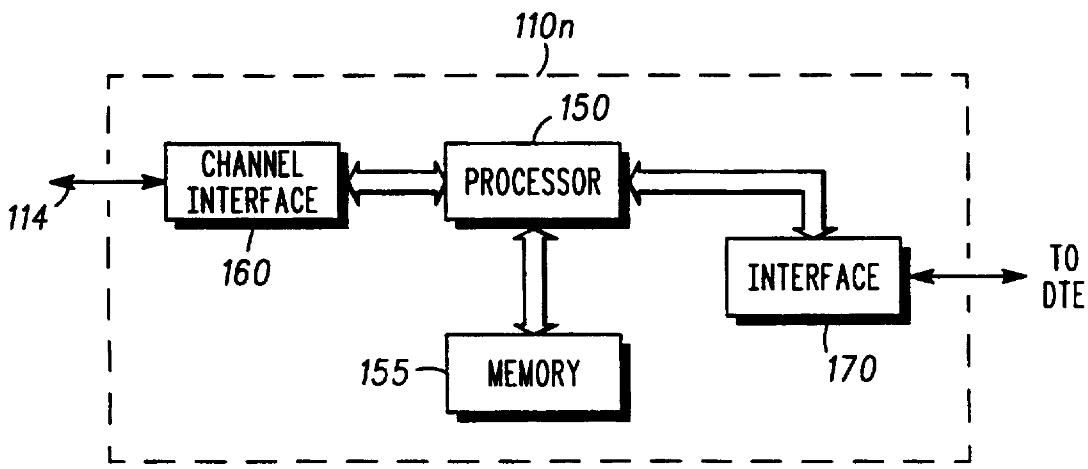

110(m+1)
110(m+2)
110n
SECONDARY STATION
SECONDARY STATION
SECONDARY STATION
100
116
NETWORK
PRIMARY STATION
101
105
115
SECONDARY STATION
SECONDARY STATION
SECONDARY STATION
SECONDARY STATION
110a
110b
110c
110(m)

FIG. 3


110n
150
CHANNEL INTERFACE
PROCESSOR
114
160
INTERFACE
TO DTE
155
MEMORY
170

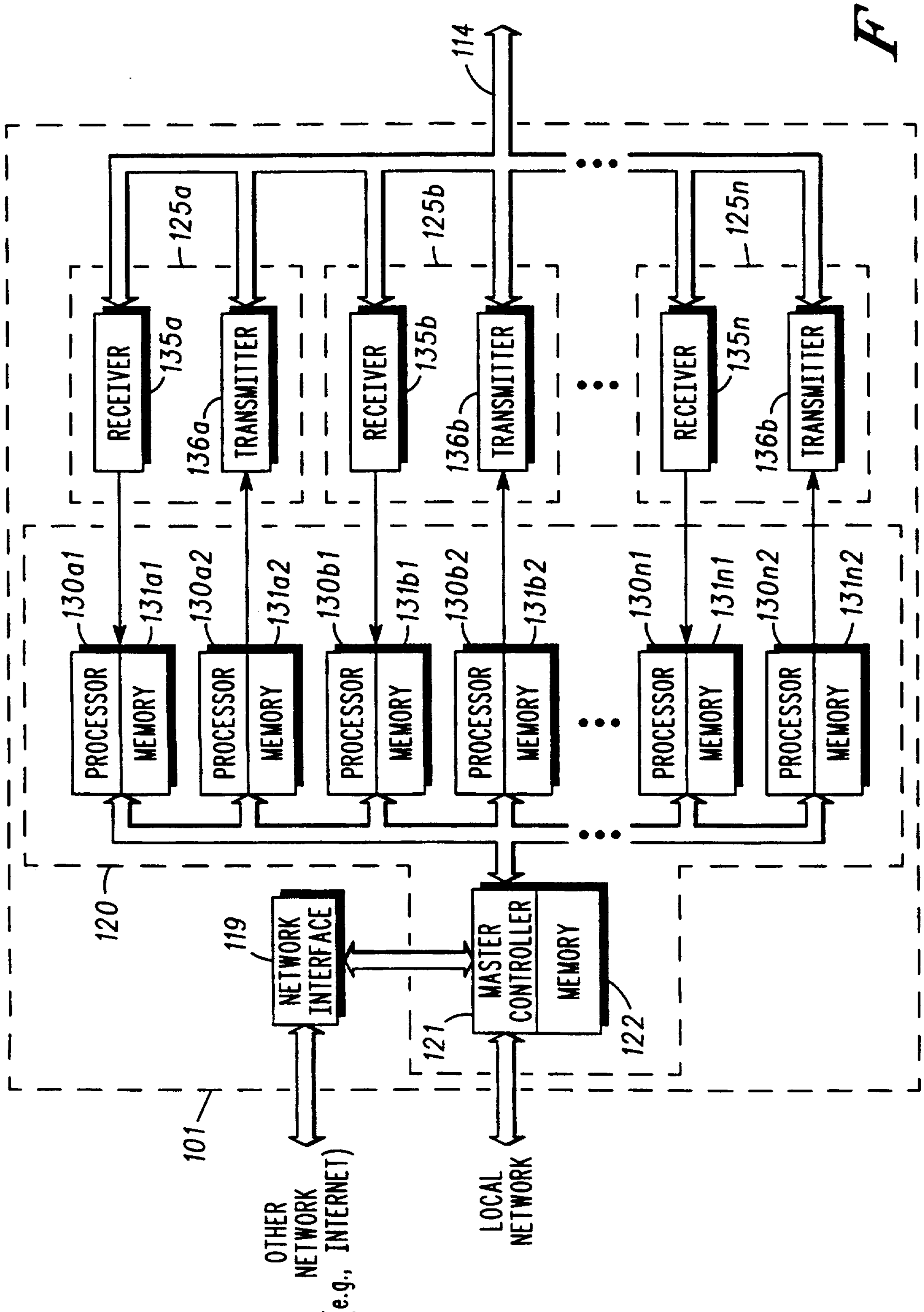

114
125a
125b
125n
RECEIVER
135a
TRANSMITTER
136a
RECEIVER
135b
TRANSMITTER
136b
RECEIVER
135n
TRANSMITTER
136b
130a1
131a1
130a2
131a2
130b1
131b1
130b2
131b2
130n1
131n1
130n2
131n2
PROCESSOR
MEMORY
120
119
NETWORK INTERFACE
MASTER CONTROLLER
MEMORY
121
122
101
OTHER NETWORK (e.g., INTERNET)
LOCAL NETWORK

FIG. 2

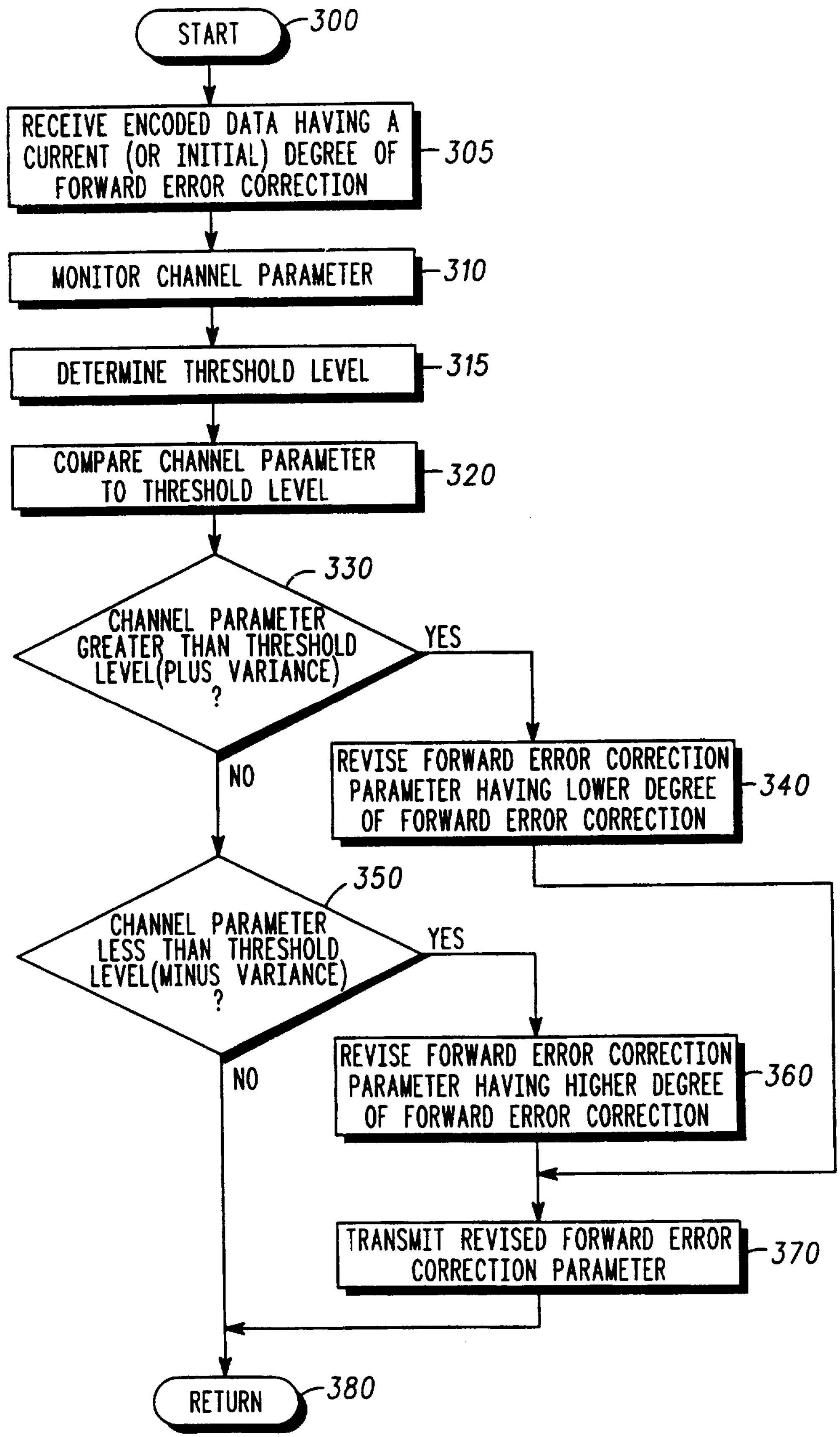

START
300
RECEIVE ENCODED DATA HAVING A CURRENT (OR INITIAL) DEGREE OF FORWARD ERROR CORRECTION
305
MONITOR CHANNEL PARAMETER
310
DETERMINE THRESHOLD LEVEL
315
COMPARE CHANNEL PARAMETER TO THRESHOLD LEVEL
320
330
CHANNEL PARAMETER GREATER THAN THRESHOLD LEVEL(PLUS VARIANCE) ?
YES
NO
REVISE FORWARD ERROR CORRECTION PARAMETER HAVING LOWER DEGREE OF FORWARD ERROR CORRECTION
340
350
CHANNEL PARAMETER LESS THAN THRESHOLD LEVEL(MINUS VARIANCE) ?
YES
NO
REVISE FORWARD ERROR CORRECTION PARAMETER HAVING HIGHER DEGREE OF FORWARD ERROR CORRECTION
360
TRANSMIT REVISED FORWARD ERROR CORRECTION PARAMETER
370
RETURN
380

FIG. 4

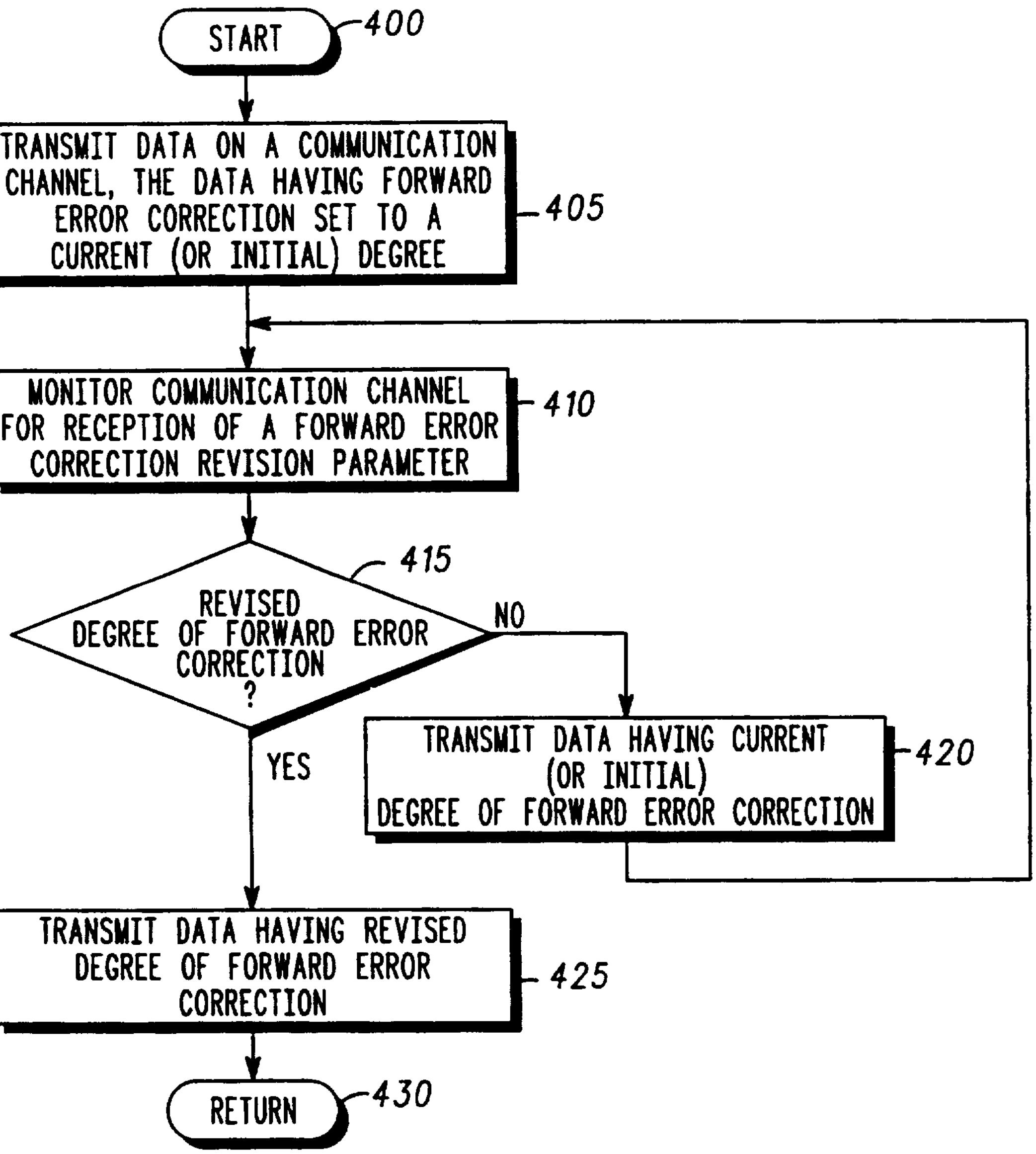

START
400
TRANSMIT DATA ON A COMMUNICATION CHANNEL, THE DATA HAVING FORWARD ERROR CORRECTION SET TO A CURRENT (OR INITIAL) DEGREE
405
MONITOR COMMUNICATION CHANNEL FOR RECEPTION OF A FORWARD ERROR CORRECTION REVISION PARAMETER
410
REVISED DEGREE OF FORWARD ERROR CORRECTION ?
415
NO
TRANSMIT DATA HAVING CURRENT (OR INITIAL) DEGREE OF FORWARD ERROR CORRECTION
420
YES
TRANSMIT DATA HAVING REVISED DEGREE OF FORWARD ERROR CORRECTION
425
RETURN
430

FIG.5

INTERNATIONAL SEARCH REPORT

International application No.
PCT/US97/04806

A. CLASSIFICATION OF SUBJECT MATTER

IPC(6) :H03M 13/00
US CL :371/41
According to International Patent Classification (IPC) or to both national classification and IPC

B. FIELDS SEARCHED

Minimum documentation searched (classification system followed by classification symbols)

U.S. : 371/41

Documentation searched other than minimum documentation to the extent that such documents are included in the fields searched

NONE

Electronic data base consulted during the international search (name of data base and, where practicable, search terms used)

NONE

C. DOCUMENTS CONSIDERED TO BE RELEVANT

Category*	Citation of document, with indication, where appropriate, of the relevant passages	Relevant to claim No.
A,P	US, 5,546,411 A (LEITCH ET AL) 13 August 1996, abstract.	1-10
A,P	US, 5,600,663 A (AYANOGLU ET AL) 04 February 1997, abstract.	1-10

☐ Further documents are listed in the continuation of Box C. ☐ See patent family annex.

* Special categories of cited documents:

"A" document defining the general state of the art which is not considered to be of particular relevance

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"L" document which may throw doubts on priority claim(s) or which is cited to establish the publication date of another citation or other special reason (as specified)

"O" document referring to an oral disclosure, use, exhibition or other means

"P" document published prior to the international filing date but later than the priority date claimed

"T" later document published after the international filing date or priority date and not in conflict with the application but cited to understand the principle or theory underlying the invention

"X" document of particular relevance; the claimed invention cannot be considered novel or cannot be considered to involve an inventive step when the document is taken alone

"Y" document of particular relevance; the claimed invention cannot be considered to involve an inventive step when the document is combined with one or more other such documents, such combination being obvious to a person skilled in the art

"&" document member of the same patent family

Date of the actual completion of the international search	Date of mailing of the international search report
10 JUNE 1997	28 JUL 1997

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Commissioner of Patents and Trademarks Box PCT Washington, D.C. 20231	STEPHEN M. BAKER
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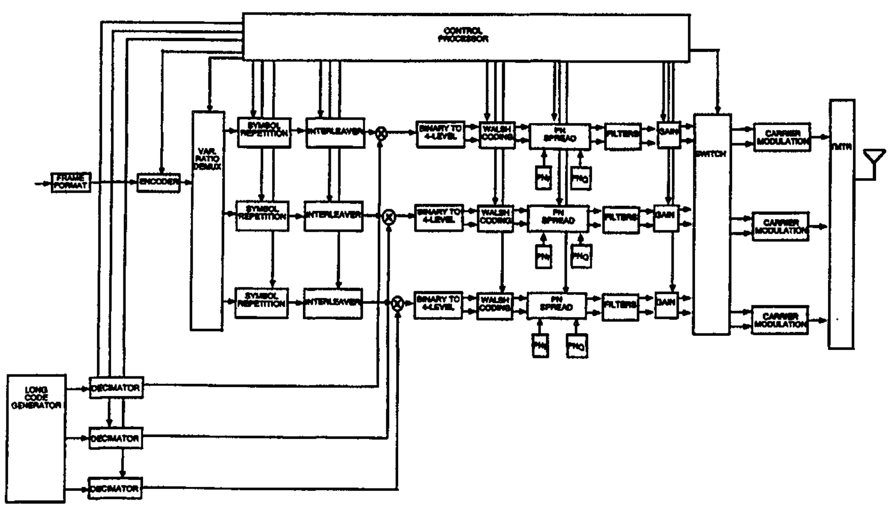
(54) Title: A METHOD OF AND APPARATUS FOR TRANSMITTING DATA IN A MULTIPLE CARRIER SYSTEM



(57) Abstract

A method of an apparatus for transmitting data in a multiple carrier system comprises encoding data and dividing the resulting encoded symbols for transmission on different frequencies. The transmitter comprises a control processor (50) for determining the capacity of each of a plurality of channels and selecting a data rate for each channel depending on the determined capacity. A plurality of transmission subsystems (56 to 72) are responsive to the control processor (50). Each transmission subsystem is associated with a respective one of the plurality of channels for scrambling encoded data with codes unique to the channel for transmission in the channel. A variable demultiplexer (56) under the control of the control processor (50) demultiplexes the encoded data into the plurality of transmission subsystems at a demultiplexing rate derived from the data rates selected for the channels by the controller. In one embodiment of the transmission subsystems, the encoded symbols are provided to a symbol repetition unit (58) which keeps the symbol rate of data to be transmitted fixed. In another embodiment, no symbol repetition is provided and variable length Walsh sequences are used to handle data rate variations.

A METHOD OF AND APPARATUS FOR TRANSMITTING DATA IN A MULTIPLE CARRIER SYSTEM

BACKGROUND OF THE INVENTION

I. Field of the Invention

The present invention relates to a method of and apparatus for transmitting data in a multiple carrier system. The present invention may be used for maximizing system throughput and increasing signal diversity by dynamically multiplexing signals onto multiple carriers in a spread spectrum communication system.

II. Description of the Related Art

It is desirable to be able to transmit data at rates which are higher than the maximum data rate of a single CDMA channel. A traditional CDMA channel (as standardized for cellular communication in the United States) is capable of carry digital data at a maximum rate of 9.6 bits per second using a 64 bit Walsh spreading function at 1.2288 MHz.

Many solutions to this problem have been proposed. One solution is to allocate multiple channels to the users and allow those users to transmit and receive data in parallel on the plurality of channels available to them. Two methods for providing multiple CDMA channels for use by a single user are described in co-pending U.S. Patent Application Serial No. 08/431,180, entitled "METHOD AND APPARATUS FOR PROVIDING VARIABLE RATE DATA IN A COMMUNICATIONS SYSTEM USING STATISTICAL MULTIPLEXING", filed April 28, 1997 and U.S. Patent Application Serial No. 08/838,240, entitled "METHOD AND APPARATUS FOR PROVIDING VARIABLE RATE DATA IN A COMMUNICATIONS SYSTEM USING NON-ORTHOGONAL OVERFLOW CHANNELS", filed April 16, 1997, both of which are assigned to the assignee of the present invention and are incorporated by reference herein. In addition, frequency diversity can be obtained by transmitting data over multiple spread spectrum channels that are separated from one another in frequency. A method and apparatus for redundantly transmitting data over multiple CDMA channels is described in U.S. Patent No. 5,166,951, entitled "HIGH CAPACITY SPREAD SPECTRUM CHANNEL", which is incorporated by reference herein.

The use of code division multiple access (CDMA) modulation techniques is one of several techniques for facilitating communications in which a large number of system users are present. Other multiple access communication system techniques, such as time division multiple access (TDMA), frequency division multiple access (FDMA) and AM modulation schemes such as amplitude companded single sideband (ACSSB) are known in the art. However, the spread spectrum modulation technique of CDMA has significant advantages over these other modulation techniques for multiple access communication systems.

The use of CDMA techniques in a multiple access communication system is disclosed in U.S. Patent No. 4,901,307, entitled "SPREAD SPECTRUM MULTIPLE ACCESS COMMUNICATION SYSTEM USING SATELLITE OR TERRESTRIAL REPEATERS", assigned to the assignee of the present invention and incorporated by reference herein. The use of CDMA techniques in a multiple access communication system is further disclosed in U.S. Patent No. 5,103,459, entitled "SYSTEM AND METHOD FOR GENERATING SIGNAL WAVEFORMS IN A CDMA CELLULAR TELEPHONE SYSTEM", assigned to the assignee of the present invention and incorporated by reference herein. Code division multiple access communications systems have been standardized in the United States in Telecommunications Industry Association Interim Standard IS-95, entitled "Mobile Station-Base Station Compatibility Standard for Dual Mode Wideband Spread Spectrum Cellular System", which is incorporated by reference herein.

The CDMA waveform by its inherent nature of being a wideband signal offers a form of frequency diversity by spreading the signal energy over a wide bandwidth. Therefore, frequency selective fading affects only a small part of the CDMA signal bandwidth. Space or path diversity on the forward/reverse link is obtained by providing multiple signal paths through simultaneous links to/from a mobile user through two or more antennas, cell sectors or cell-sites. Furthermore, path diversity may be obtained by exploiting the multipath environment through spread spectrum processing by allowing a signal arriving with different propagation delays to be received and processed separately. Examples of the utilization of path diversity are illustrated in co-pending U.S. Patent No. 5,101,501 entitled "SOFT HANDOFF IN A CDMA CELLULAR TELEPHONE SYSTEM", and U.S. Patent No. 5,109,390 entitled "DIVERSITY RECEIVER IN A CDMA CELLULAR TELEPHONE SYSTEM", both assigned to the assignee of the present invention and incorporated by reference herein.

FIG. 1 illustrates a transmission scheme for a multiple-carrier code division multiple access (CDMA) system in which each carrier carries a fixed fraction of the transmitted data. Variable rate frame of information bits are provided to encoder **2** which encodes the bits in accordance with a convolutional encoding format. The encoded symbols are provided to symbol repetition means **4**. Symbol repetition means **4** repeats the encoded symbols so as to provide a fixed rate of symbols out of symbol repetition means **4**, regardless of the rate of the information bits.

The repeated symbols are provided to block interleaver **6** which rearranges the sequence in which the symbols are to be transmitted. The interleaving process, coupled with the forward error correction, provides time diversity which aids in the reception and error recovery of the transmitted signal in the face of burst errors. The interleaved symbols are provided to data scrambler **12**. Data scrambler **12** multiplies each interleaved symbol by +1 or -1 according to a pseudonoise (PN) sequence. The pseudonoise sequence is provided by passing a long PN sequence generated by long code generator **8** at the chip rate through decimator **10** which selectively provides a subset of the chips of the long code sequence at the rate of the interleaved symbol stream.

The data from data scrambler **12** is provided to demultiplexer (DEMUX) **14**. Demultiplexer **14** divides the data stream into three equal sub-streams. The first sub-stream is provided to transmission subsystem **15a**, the second sub-stream to transmission subsystem **15b** and the third sub-stream to transmission subsystem **15c**. The subframes are provided to serial-to-parallel converters (BINARY TO 4 LEVEL) **16a-16c**. The outputs of serial to parallel converters **16a-16c** are quaternary symbols (2bits/symbol) to be transmitted in a QPSK modulation format

The signals from serial-to-parallel converters **16a-16c** are provided to Walsh coders **18a-18c**. In Walsh coders **18a-18c**, the signal from each converter **16a-16k** is multiplied by a Walsh sequence consisting of ±1 values. The Walsh coded data is provided to QPSK spreaders **20a-20c**, which spread the data in accordance with two short PN sequences. The short PN sequence spread signals are provided to amplifiers **22a-22b** which amplify the signals in accordance with a gain factor.

The system described above suffers from a plurality of drawbacks. First, because the data is to be provided in equal sub-streams on each of the carriers, the available numerology is limited to frames with a number of code symbols that will divide evenly by a factor of three. Table 1 below

illustrates the limited number of possible rate sets which are available using the transmission system illustrated in FIG. 1.

Walsh Function (QPSK Symbol) Rate [sps]	Number of Walsh Functions per 20ms		Length of Walsh Function [chips]	Symbol Rate [sps] (After Repetition)	Number of Symbols per 20 ms.	
1228800	24576	$3*(2^{13})$	1	2457600	49152	$3*(2^{14})$
614400	12288	$3*(2^{12})$	2	1228800	24576	$3*(2^{13})$
307200	6144	$3*(2^{11})$	4	614400	12288	$3*(2^{12})$
153600	3072	$3*(2^{10})$	8	307200	61444	$3*(2^{11})$
76800	1536	$3*(2^{9})$	16	153600	3072	$3*(2^{10})$
38400	768	$3*(2^{8})$	32	76800	1536	$3*(2^{9})$
19200	384	$3*(2^{7})$	64	38400	768	$3*(2^{8})$
9600	192	$3*(2^{6})$	128	19200	384	$3*(2^{7})$
4800	96	$3*(2^{5})$	256	9600	192	$3*(2^{6})$
2400	48	$3*(2^{4})$	512	4800	96	$3*(2^{5})$
1200	24	$3*(2^{3})$	1024	2400	48	$3*(2^{4})$
600	12	$3*(2^{2})$	2048	1200	24	$3*(2^{3})$
300	6	$3*(2^{1})$	4096	600	12	$3*(2^{2})$
150	3	$3*(2^{0})$	8192	300	6	$3*(2^{1})$

Table 1

As illustrated in Table 1, because the symbols are evenly distributed to the three carriers, the total data rate is limited by the carrier with the least power available or requiring the highest SNR. That is the total data rate is equalt to three times te data rate of the "worst" link (here the worst means the one requiring the highest SNR or having the least power avaialble). this reduces the system throughput, because the worst link's rate is always chosen as the common rate for all three carriers, which results in under utilization of the channel resource on the two better links.

Second, frequency dependent fading can severely affect one of the frequencies while having a limited effect on the remaining frequencies. This implementation is inflexible and does not allow transmission of a frame to be provided in a way that reduces the effects of the poor channel. Third, because of frequency dependent fading, the fading will typically always affectthe same groups of symbolsof each frame. Fourth, were the

implementation to be superimposed on a speech transmission system there is no good way to balance the loads carried on the different frequencies on a frame by frame basis in the face of variable speech activities in each frame. This results in loss in total system throughput. And fifth, for a system with only three frequency channels, with the implementation described, there is no method of separating the speech and data so as to provide the data on one frequency or set of frequencies and the speech on a different frequency or set of frequencies. This results in a loss of system throughput as mentioned above.

Therefore, there is a need felt for an improved multi-carrier CDMA communication system which offers greater flexibility in numerology and load balancing, better resolution in data rates supported, and which offers superior performance in the face of frequency dependent fading and uneven loading.

SUMMARY OF THE INVENTION

In one aspect the invention provides a transmitter for transmitting data at a data rate in a plurality of channels each having a capacity less than the data rate, the transmitter comprising: a controller for determining the capacity of each of a plurality channels and selecting a data rate for each channel depending on the determined capacity; a plurality of transmission subsystems responsive to the controller and each associated with a respective one of the plurality of channels for scrambling encoded data with codes unique to the channel for transmission in the channel; and a variable demultilplexer responsive to the controller for demultiplexing the encoded data into the plurality of transmission subsystems at a demultiplexing rate derived from the data rates selected for the channels by the controller.

In another aspect the invention provides a receiver comprising: a receiving circuit for receiving signals simultaneously in a plurality of channels each of which signals define scrambled encoded symbols which together represent data from a common origin; a controller for determining a symbol rate for the signals in each channel; a plurality of receiving subsystems responsive to the controller and each associated with a respective one of the plurality of channels for descrambling encoded symbols with codes unique to the channel to enable the data to be extracted therefrom; and a variable multilplexer responsive to the controller for multiplexing the data from the plurality of receiving subsystems at a

multiplexing rate derived from the symbol rates determined for the channels by the controller onto an output.

In a further aspect the invention provides a wireless transmitter, comprising: encoder for receive a set of information bits and encoding said information bits to provide a set of code symbols; and a transmission subsystem for receiving said code symbols and for providing a subset of said code symbols on a first carrier frequency and the remaining symbols on at least one additional carrier frequency.

The invention also provides a method of transmitting data at a data rate in a plurality of channels each having a capacity less than the data rate, the method comprising: determining the capacity of each of a plurality channels and selecting a data rate for each channel depending on the determined capacity; scrambling encoded data with codes unique to the channel for transmission in the channel; and demultiplexing the encoded data into the plurality of channels at a demultiplexing rate derived from the data rates selected for the channels by the controller.

The invention further provides a method of receiving data, the method comprising: receiving signals simultaneously in a plurality of channels each of which signals define scrambled encoded symbols which together represent data from a common origin; determining a symbol rate for the signals in each channel; descrambling encoded symbols in each channel with codes unique to the channel to enable the data to be extracted therefrom; and multiplexing the descrambled data from the plurality of channels at a multiplexing rate derived from the symbol rates determined for the channels.

To better utilize the channel resource, it's necessary to be able to transmit a different data rate on each carrier according to the channel condition and the available power on each channel. One way of doing this is by changing the ratio of the inverse-multiplexing on to each of the carriers. Instead of distributing the symbols with a ratio of 1:1:1, a more arbitrary ratio can be used together with different repetition schemes as long as the resulted symbol rate on each carrier is a factor of some Walsh function rate. Walsh function rate can be 1228800, 614400, 307200,..., 75 for Walsh function length from 1 to 16384.

Given the Walsh function length, if the symbol rate is lower than the Walsh function rate, symbol repetition is used to "match" the rate. The repetition factor can be any number, integer or fractional. It will be understood by one skilled in the art that when repetition is present, the total transmit power can be proportionately reduced to keep the code symbol

energy constant. The Walsh function length may or may not be the same on the three carriers, depending on whether we need to save code channels. For example, if the supportable code symbol rate on the three channels are 153600 sps, 30720 sps and 102400 sps (for rate 1/2 coding, these correspond to data rates of 76.8 kbps, 15.36 kbps and 51.2 kbps, respectively - the total data rate is 143.36 kbps), then the inverse-multiplexing ratio will be 15:3:10.

If a Walsh function of length 8 is used for all three channels (assuming QPSK modulation with a QPSK symbol rate of 153.6 Ksps), then each code symbol is transmitted twice, 10 times, and three times on the three channels, respectively. Additional time diversity can be obtained if the repeated symbols are further interleaved. In an alternative embodiment, different Walsh function lengths are used. For example, Walsh functions for the three channels in the example of above of length 16, 16 and 8 respectively can be used, with each code symbol transmitted once on the first channel, five times on the second, and three times on the third.

The above approach does not affect the encoder since it has to be able to handle the highest data rate anyway. All that is changed is the number of data octets at the encoder input. However, this approach does have an impact on the implementation of the interleaver because the interleaver will have many possible sizes (in terms of number of symbols) if all combinations of data rates on the three channels are allowed. One alternative to the above approach which mitigates this problem is to inverse-multiplex the code symbols out of the encoder to the three carriers directly and perform interleaving of repeated code symbols on each channel separately. This simplifies the numerology and reduces the number of possible interleaver sizes on each channel.

BRIEF DESCRIPTION OF THE DRAWINGS

Further features, objects, and advantages of the present invention will become more apparent from the detailed description set forth below of embodiments of the invention when taken in conjunction with the drawings in which like reference characters identify correspondingly throughout and wherein:

FIG. 1 is a block diagram illustrating a multiple frequency CDMA communication system with fixed rates and carriers;

FIG. 2 is a block diagram illustrating a transmission system embodying the present invention;

FIG. 3 is a block diagram illustrating a receiver system embodying the present invention; and

FIG. 4 is a table of code channel Walsh symbols in a traditional IS-95 CDMA communication system.

DETAILED DESCRIPTION OF THE PREFERRED EMBODIMENTS

Referring to FIG. 2, which is a block diagram illustrating a transmission system embodying the present invention, the first operation to be performed is to determine the amount of data which can be supported on each of the carriers. Three such carriers are illustrated in FIG. 2, though one skilled in the art will realize that the present invention is easily extended to any number of carriers. Control processor **50** based on a set of factors such as the loading on each of the carriers, the amount of data queued for transmission to the mobile station and the priority of the information to be transmitted to the mobile station determines the rate of data transmission on each of the carriers.

After having selected the data rate to be transmitted on each of the carriers, control processor **50** selects a modulation format that is capable of transmitting data at the selected rate. In the exemplary embodiment, different length Walsh sequences are used to modulate the data depending on the rate of the data to be transmitted. The use of different length Walsh sequences selected to modulate the data depending on the rate of the data to be transmitted is described in detail in co-pending U.S. Patent Application Serial No. 08/654,443, filed May 28, 1996, entitled "HIGH RATE DATA WIRELESS COMMUNICATION SYSTEM", which is assigned to the assignee of the present invention and incorporated by reference herein. In an alternative embodiment, the high rate data can be supported by bundling of CDMA channels as described in the aforementioned Patent Applications Serial Nos. 08/431,180 and 08/838,240.

Once the rates which will be supported on each of the carriers is selected then control processor **50** calculates an inverse multiplexing ratio that will determine the amount of each transmission that will be carried on each of the carriers. For example, if the supportable code symbol rate on the three channels are 153600 sps, 30720 sps and 102400 sps (for rate 1/2 coding, these correspond to data rates of 76.8 kbps, 15.36 kbps and 51.2 kbps, respectively - the total data rate is 143.36 kbps), then the inverse-multiplexing ratio will be 15:3:10.

In the exemplary embodiment, frames of information bits are provided to frame formatter **52**. In the exemplary embodiment, formatter **52** generates and appends to the frame a set of cyclic redundancy check (CRC) bits. In addition, formatter **52** appends a predetermined set of tail bits. The implementation and design of frame formatters are well known in the art, an example of a typical frame formatter is described in detail in U.S. Patent No. 5,600,754, entitled "METHOD AND SYSTEM FOR THE ARRANGEMENT OF VOCODER DATA FOR THE MASKING OF TRANSMISSION CHANNEL INDUCED ERRORS", which is assigned to the assignee of the present invention and incorporated by reference herein.

The formatted data is provided to encoder **54**. In the exemplary embodiment, encoder **54** is a convolutional encoder, though the present invention can be extended to other forms of encoding. A signal from control processor **50** indicates to encoder **54** the number of bits to be encoded in this transmission cycle. In the exemplary embodiment, encoder **54** is a rate 1/4 convolutional encoder with a constraint length of 9. It should be noted that because of the additional flexibility provided by the present invention, essentially any encoding format can be used.

The encoded symbols from encoder **54** are provided to variable ratio de-multiplexer **56**. Variable ratio de-multiplexer **56** provides the encoded symbols to a set of outputs based on a symbol output signal provided by control processor **50**. In the exemplary embodiment, there are three carrier frequencies and control processor **50** provides a signal indicative of the number of encoded symbols to be provided on each of the three outputs. As one skilled in the art will appreciate, the present invention is easily extended to an arbitrary number of frequencies.

The encoded symbols provided on each of the outputs of variable ratio de-multiplexer **56** are provided to a corresponding symbol repetition means **58a-58c**. Symbol repetition means **58a-58c** generate repeated versions of the encoded symbols so that the resultant symbol rate matches with the rate of data supported on that carrier and the in particular matches Walsh function rate used on that carrier. The implementation of repetition generators **58a-58c** is known in the art and an example of such is described in detail in U.S. Patent No. 5,629,955, entitled "Variable Response Filter", which is assigned to the assignee of the present invention and incorporated by reference herein. Control processor **50** provides a separate signal to each repetition generator **58a-58c** indicating the rate of symbols on each carrier or alternatively the amount of repetition to be provided on each carrier. In response to the signal from control processor **50**, repetition means **58a-58c**

generate the requisite numbers of repeated symbols to provide the designated symbol rates. It should be noted that in the preferred embodiment, the amount of repetition is not limited to integer number wherein all symbols are repeated the same number of times. A method for providing non-integer repetition is described in detail in co-pending U.S. Patent Application Serial No. 08/886,815, filed March 26, 1997, entitled "METHOD AND APPARATUS FOR TRANSMITTING HIGH SPEED DATA IN A SPREAD SPECTRUM COMMUNICATIONS SYSTEM", which is assigned to the assignee of the present invention and incorporated by reference herein.

The symbols from repetition generators **58a-58c** are provided to a corresponding one of interleavers **60a-60c** which reorders the repeated symbols in accordance with a predetermined interleaver format. Control processor **50** provides an interleaving format signal to each of interleavers **60a-60c** which indicates one of a predetermined set of interleaving formats. In the exemplary embodiment, the interleaving format is selected from a predetermined set of bit reversal interleaving formats.

The reordered symbols from interleavers **60a-60c** are provided to data scramblers **62a-62c**. Each of data scramblers **62a-62c** changes the sign of the data in accordance with a pseudonoise (PN) sequence. Each PN sequence is provided by passing a long PN code generated by long code or PN generator **82** at the chip rate through a decimator **84a-84c**, which selectively provides ones of the spreading symbols to provide a PN sequence at a rate no higher than that provided by PN generator **82**. Because the symbol rate on each carrier may be different from one another, the decimation rate of decimators **84a-84c** may be different. Decimators **84a-84c** are sample and hold circuits which sample the PN sequence out of PN generator **82** and continue to output that value for a predetermined period. The implementation of PN generator **82** and decimators **84a-84c** are well known in the art and are described in detail in the aforementioned U.S. Patent No. 5,103,459. Data scramblers **62a-62c** exclusively-OR the binary symbols from interleavers **60a-60c** with the decimated pseudonoise binary sequences from decimators **84a-84c**.

The binary scrambled symbol sequences are provided to serial to parallel converters (BINARY TO 4-LEVEL) **64a-64c**. Two binary symbols provided to converters **64a-64c** are mapped to a quaternary constellation with values (±1, ±1). The constellation values are provided on two outputs from converters **62a-62c**. The symbol streams from converters **64a-64c** are separately provided to Walsh spreaders **66a-66c**.

There are many methods of providing high speed data in a code division multiple access communication system. In the preferred embodiment, the Walsh sequence length is varied in accordance with the rate of the data to be modulated. Shorter Walsh sequences are used to modulate higher speed data and longer Walsh sequences are used to modulate lower rate data. For example, a 64 bit Walsh sequence can be used to transmit data at 19.2 Ksps. However, a 32 bit Walsh sequence can be used to modulate data at 38.4 Ksps.

A system describing variable length Walsh sequence modulation is described in detail in co-pending U.S. Patent Application Serial No. 08/724,281, entitled "HIGH DATA RATE SUPPLEMENTAL CHANNEL FOR CDMA TELECOMMUNICATIONS SYSTEM", filed January 15, 1997 and incorporated by reference herein. The length of the Walsh sequences used to modulate the data depend on the rate of the rate of the data to be transmitted. FIG. 4 illustrates the Walsh functions in a traditional IS-95 CDMA system.

In the preferred embodiment of the invention, the number of Walsh channels allocated for the high-rate data can be any value 2^N where N = {2, 3, 4, 5, 6}. The Walsh codes used by Walsh coders **66a-66c** are $64/2^N$ symbols long, rather than the 64 symbols used with the IS-95 Walsh codes. In order for the high-rate channel to be orthogonal to the other code channels with 64-symbol Walsh codes, 2^N of the possible 64 quaternary-phase channels with 64-symbol Walsh are eliminated from use. Table I provides a list of the possible Walsh codes for each value of N and the corresponding sets of allocated 64-symbol Walsh codes.

N	$Walsh_i$	Allocated 64-Symbol Walsh Codes
2	+,+,+,+,+,+,+,+,+,+,+,+,+,+,+,+	0, 16, 32, 48
	+,−,+,−,+,−,+,−,+,−,+,−,+,−,+,−	1, 17, 33, 49
	+,+,−,−,+,+,−,−,+,+,−,−,+,+,−,−	2, 18, 34, 50
	+,−,−,+,+,−,−,+,+,−,−,+,+,−,−,+	3, 19, 35, 51
	+,+,+,+,−,−,−,−,+,+,+,+,−,−,−,−	4, 20, 36, 52
	+,−,+,−,−,+,−,+,+,−,+,−,−,+,−,+	5, 21, 37, 53
	+,+,−,−,−,−,+,+,+,+,−,−,−,−,+,+	6, 22, 38, 54
	+,−,−,+,−,+,+,−,+,−,−,+,−,+,+,−	7, 23, 39, 55
	+,+,+,+,+,+,+,+,−,−,−,−,−,−,−,−	8, 24, 40, 56
	+,−,+,−,+,−,+,−,−,+,−,+,−,+,−,+	9, 25, 41, 57
	+,+,−,−,+,+,−,−,−,−,+,+,−,−,+,+	10, 26, 42, 58
	+,−,−,+,+,−,−,+,−,+,+,−,−,+,+,−	11, 27, 43, 59
	+,+,+,+,−,−,−,−,−,−,−,−,+,+,+,+	12, 28, 44, 60
	+,−,+,−,−,+,−,+,−,+,−,+,+,−,+,−	13, 29, 45, 61
	+,+,−,−,−,−,+,+,−,−,+,+,+,+,−,−	14, 30, 46, 62
	+,−,−,+,−,+,+,−,−,+,+,−,+,−,−,+	15, 31, 47, 63
3	+,+,+,+,+,+,+,+	0, 8, 16, 24, 32, 40, 48, 56
	+,−,+,−,+,−,+,−	1, 9, 17, 25, 33, 41, 49, 57
	+,+,−,−,+,+,−,−	2, 10, 18, 26, 34, 42, 50, 58
	+,−,−,+,+,−,−,+	3, 11, 19, 27, 35, 43, 51, 59
	+,+,+,+,−,−,−,−	4, 12, 20, 28, 36, 44, 52, 60
	+,−,+,−,−,+,−,+	5, 13, 21, 29, 37, 45, 53, 61
	+,+,−,−,−,−,+,+	6, 14, 22, 30, 38, 46, 54, 62
	+,−,−,+,−,+,+,−	7, 15, 23, 31, 39, 47, 55, 63
4	+,+,+,+	0, 4, 8,…,60
	+,−,+,−	1, 5, 9, …,61
	+,+,−,−	2, 6, 10, …,62
	+,−,−,+	3, 7, 11, …,63
5	+,+	0, 2, 4,…,62
	+,−	1, 3, 5, …,63
6	+	0, 1, 2,…,63

Table I.

The + and − indicate a positive or negative integer value, where the preferred integer is 1. As is apparent, the number of Walsh symbols in each Walsh code varies as N varies, and in all instances is less than the number

of symbols in the IS-95 Walsh channel codes. Regardless of the length of the Walsh code, in the described embodiment of the invention the symbols are applied at a rate of 1.2288 Megachips per second (Mcps). Thus, shorter length Walsh codes are repeated more often. Control processor **50** provides a signal to Walsh coding elements **66a-66c** which indicates the Walsh sequence to be used to spread the data.

Alternative methods for transmitting high rate data in CDMA communication system also include methods generally referred to as channel bundling techniques. The present invention is equally applicable to the channel bundling methods for providing high speed data in a CDMA communciation system. One method of providing channel bundled data is to provide a plurality of Walsh channels for use by a signal user. This method is described in detail in the aforementioned U.S. Patent Application Serial No. 08/739,482. An alternative channel bundling technique is to provide the user with use of one Walsh code channel but to differentiate the signals from one another by means of different scrambling signals as described in detail in co-pending U.S. Patent Application Serial No. 08/838,240.

The Walsh spread data is provided to PN spreaders **68a-68c**, which apply a short PN sequence spreading on the output signals. In the exemplary embodiment, the PN spreading is performed by means of a complex multiplication as described in detail in the aforementioned co-pending U.S. Patent Application Serial No. 08/784,281. Data channels D_I and D_Q are complex multiplied, as the first real and imaginary terms respectively, with spreading codes PN_I and PN_Q, as the second real and imaginary terms respectively, yielding in-phase (or real) term X_I and quadrature-phase (or imaginary) term X_Q. Spreading codes PN_I and PN_Q are generated by spreading code generators **67** and **69**. Spreading codes PN_I and PN_Q are applied at 1.2288 Mcps. Equation (1) illustrates the complex multiplication performed.

$$(X_I + jX_Q) = (D_I + jD_Q)(PN_I + jPN_Q) \quad (1)$$

In-phase term X_I is then low-pass filtered to a 1.2288 MHz bandwidth (not shown) and upconverted by multiplication with in-phase carrier $COS(\omega_c t)$. Similarly, quadrature-phase term X_Q is low-pass filtered to a 1.2288 MHz bandwidth (not shown) and upconverted by multiplication with quadrature-phase carrier $SIN(\omega_c t)$. The upconverted X_I and X_Q terms are summed yielding forward link signal s(t). The complex multiplication

allows quadrature-phase channel set to remain orthogonal to the in-phase channel set and therefore to be provided without adding additional interference to the other channels transmitted over the same path with perfect receiver phase recovery.

The PN spread data is, then, provided to filters **70a-70c** which spectrally shape the signals for transmission. The filtered signals are provided to gain multipliers **72a-72c**, which amplify the signals for each carrier. The gain factor is supplied to gain elements **72a-72c** by control processor **50**. In the exemplary embodiment, control processor **50** selects the gain factor for each carrier in accordance with the channel condition and the rate of the information data to be transmitted on that carrier. As is known by one skilled in the art, data that is transmitted with repetition can be transmitted with lower symbol energy than data without repetition.

The amplified signals are provided to an optional switch **74**. Switch **74** provides the additional flexibility of channel hopping the data signals onto different carriers. Typically, switch **74** is only used when the number of carriers actually used to transmit the signal is smaller that the total number of possible carriers (3 in the present example).

The data is passed by switch **74** to carrier modulators **76a-76c.** Each of carrier modulators **76a-76c** upconvert the data to a different predetermined frequency. The upconverted signals are provided to transmitter **78** where they are combined with other similarly processed signals, filtered and amplified for transmission through antenna **80**. In the exemplary embodiment, the amplified frequency upon which each of the signals are transmitted varies with time. This provides additional frequency diversity for the transmitted signals. For example a signal that is currently being transmitted through carrier modulator **76a** will at predetermined time interval be switched so as to be transmitted on a different frequency through carrier modulators **76b** or **76c**. In accordance with a signal from control processor **50**, switch **74** directs an amplified input signal from gain multiplier **72a-72c** to an appropriate carrier modulator **76a-76c**.

Turning to FIG. 3, a receiver system embodying the present invention is illustrated. The signal received at antenna **100** is passed to receiver (RCVR) **102**, which amplifies and filters the signal before providing it to switch **104**. The data is provided through switch **104** to an appropriate carrier demodulator **106a-106c**. It will be understood by one skilled in the art that although the receiver structure is described for the reception of a signal transmitted on three frequencies, the present invention can easily be

extended to an arbitrary number of frequencies consecutive to one another or not.

When the carriers on which the data is transmitted are rotated or hopped to provide additional frequency diversity, switch **104** provides the received signal to a selected carrier demodulator **106a-106c** in response to a control signal from control processor **125**. When the carrier frequencies are not hopped or rotated, then switch **104** is unnecessary. Each of carrier demodulators **106a-106c** Quaternary Phase Shift Keying (QPSK) demodulate the received signal to baseband using a different downconversion frequency to provide a separate I and Q baseband signals.

The downconverted signals from each of carrier demodulators **106a-106c** are provided to a corresponding PN despreader **108a-108c** which removes the short code spreading from the downconverted data. The I and Q signals are despread by complex multiplication with a pair of short PN code. The PN despread data is provided to Walsh demodulators **110a-110c**, which uncover the data in accordance with the assigned code channel sequences. In the exemplary embodiment, Walsh functions are used in the generation and reception of the CDMA signals but other forms of code channel generation are equally applicable. Control processor **125** provides a signal to Walsh demodulators **110a-110c** indicating the Walsh sequences to be used to uncover the data.

The Walsh despread symbols are provided to parallel-to-serial converters (4-LEVEL TO BINARY) **112a-112c**, which map the 2-dimensional signal into a 1-dimensional signal. The symbols are then provided to descramblers **114a-114c**. Descramblers **114a-114c** descramble the data in accordance with a decimated long code sequence generated as described with respect to the decimated long code sequences used to scramble the data in FIG. 2.

The descrambled data is provided to de-interleavers (DE-INT) **116a-116c**. De-interleavers **116a-116c** reorder the symbols in accordance with selected de-interleaver formats that are provided by control processor **125**. In the exemplary embodiment, control processor **125** provides a signal indicative of the size of the deinterleaver and the scheme of de-interleaving to each of de-interleavers **116a-116c**. In the exemplary embodiment, the de-interleaving scheme is selected from a predetermined set of bit reversal de-interleaving schemes.

The de-interleaved symbols are then provided to symbol combiners **118a-118c** which coherently combine those repeatedly transmitted symbols. The combined symbols (soft decisions) are then provided to variable ratio

multiplexer **120** which reassembles the data stream and provides the reassembled data stream to decoder **122**. In the exemplary embodiment decoder **122** is a maximum likelihood decoder, the implementation of which is well known in the art. In the exemplary embodiment, decoder **122** contains a buffer (not shown) which waits until an entire frame of data has been provided to it before beginning the decoding process. The decoded frame is provided to CRC check means **124** which determines whether the CRC bits check and if so provides them to the user otherwise an erasure is declared.

Having thus described the invention by reference to a preferred embodiment it is to be well understood that the embodiment in question is exemplary only and that modifications and variations such as will occur to those possessed of appropriate knowledge and skills may be made without departure from the spirit and scope of the invention as set forth in the appended claims and equivalents thereof.

I CLAIM:

CLAIMS

1. A transmitter for transmitting data at a data rate in a plurality of channels each having a capacity less than the data rate, the transmitter comprising;

a controller for determining the capacity of each of a plurality channels and selecting a data rate for each channel depending on the determined capacity;

a plurality of transmission subsystems responsive to the controller and each associated with a respective one of the plurality of channels for scrambling encoded data with codes unique to the channel for transmission in the channel; and

a variable demultilplexer responsive to the controller for demultiplexing the encoded data into the plurality of transmission subsystems at a demultiplexing rate derived from the data rates selected for the channels by the controller.

2. A transmitter as claimed in claim 1, further comprising an encoder for generating the encoded data from frames of data input thereto.

3. A transmitter as claimed in claim 1 or 2, wherein each transmission subsystem comprises a symbol repetition unit for repeating symbols to output the same at a rate corresponding to the rate selected for the channel by the controller.

4. A transmitter as claimed in claim 3, wherein each transmission subsystem comprises an interleaving unit for reordering the repeated symbols depending on an interleaving format determined by the controller.

5. A transmitter as claimed in claim 4, further comprising a long code generator for generating a respective long code for each channel; and

in each transmission subsystem, a scrambler for scrambling the reordered symbols using the respective code for the channel.

6. A transmitter as claimed in claim 5, wherein the long code generator comprises for each channel a decimator unit for decimating a generated long code at a decimation rate determined by the controller so as to produce the respective long codes for each channel.

7. A transmitter as claimed in claim 6, further comprising variable coding units in each transmission subsystem for modulating the scrambled symbols from the scrambler.

8. A transmitter as claimed in claim 7, wherein the coding units are arranged to modulate the scrambled symbols with a respective walsh code.

9. A transmitter as claimed in claim 7 or 8, further comprising a pseudo noise spreader in each channel for spreading the modulated symbols.

10. A transmitter as claimed in any preceding claim, further comprising:

a switch; and

a plurality of carrier modulators, wherein the switch is responsive to the controller for switching the scrambled data from the plurality of transmission subsystems between the plural carrier modulators for modulation of the signals onto different carriers at different times.

11. A receiver comprising:

a receiving circuit for receiving signals simultaneously in a plurality of channels each of which signals define scrambled encoded symbols which together represent data from a common origin;

a controller for determining a symbol rate for the signals in each channel;

a plurality of receiving subsystems responsive to the controller and each associated with a respective one of the plurality of channels for descrambling encoded symbols with codes unique to the channel to enable the data to be extracted therefrom; and

a variable multilplexer responsive to the controller for multiplexing the data from the plurality of receiving subsystems at a multiplexing rate derived from the symbol rates determined for the channels by the controller onto an output.

12. A receiver as claimed in claim 11, further comprising an decoder for decoding the encoded data output from the multiplexer into frames of data.

13. A receiver as claimed in claim 11 or 12, further comprising a pseudo noise despreader in each channel for despreading the scrambled encoded symbols.

14. A receiver as claimed in claim 13, further comprising variable decoding units in each receiving subsystem for demodulating the despread symbols from the despreader.

15. A receiver as claimed in claim 14, wherein the decoding units are arranged to demodulate the despread symbols with a respective walsh code.

16. A receiver as claimed in claim 15, further comprising, in each receiving subsystem, a descrambler for descrambling the despread symbols using a respective long code for the channel.

17. A receiver as claimed in claim 16, wherein each receiving subsystem comprises an deinterleaving unit for reordering the repeated symbols depending on an interleaving format determined by the controller.

18. A receiver as claimed in claim 17, wherein each receiving subsystem comprises a symbol combiner for combining symbols to output the same to the demultiplexer at a rate corresponding to the rate determined for the channel by the controller.

19. A receiver as claimed in any of claims 11 to 18, further comprising:

a switch; and

a plurality of carrier demodulators, wherein the switch is responsive to the controller for switching the received signals between the plural carrier demodulators for demodulation of the signals into different receiving subsystems at different times.

20. A wireless transmitter, comprising:

encoder for receive a set of information bits and encoding said information bits to provide a set of code symbols; and

transmission subsystem for receiving said code symbols and for providing a subset of said code symbols on a first carrier frequency and the remaining symbols on at least one additional carrier frequency.

21. A method of transmitting data at a data rate in a plurality of channels each having a capacity less than the data rate, the method comprising;

determining the capacity of each of a plurality channels and selecting a data rate for each channel depending on the determined capacity;

scrambling encoded data with codes unique to the channel for transmission in the channel; and

demultiplexing the encoded data into the plurality of channels at a demultiplexing rate derived from the data rates selected for the channels by the controller.

22. A method as claimed in claim 21, further comprising an encoder for generating the encoded data from frames of data input thereto.

23. A method as claimed in claim 21 or 22, further comprising repeating symbols for each channel to output the same at a rate corresponding to the rate selected for the channel.

24. A method as claimed in claim 23, further comprising reordering the repeated symbols depending on a determined interleaving format.

25. A method as claimed in claim 24, further comprising

generating a respective long code for each channel; and

scrambling the reordered symbols in each transmission subsystem using the respective code for the channel.

26. A method as claimed in claim 25, wherein the long code is generated for each by decimating a generated long code at a determined decimation rate for each channel.

27. A method as claimed in claim 26, further comprising for modulating the scrambled symbols with a code.

28. A method as claimed in claim 27, the scrambled symbols are modulated with a respective walsh code.

29. A method as claimed in claim 27 or 28, further comprising spreading the modulated symbols with pseudo noise.

30. A method as claimed in any of claims 21 to 29, further comprising modulating the scrambled data onto different carriers at different times.

31. A method of receiving data, the method comprising:

receiving signals simultaneously in a plurality of channels each of which signals define scrambled encoded symbols which together represent data from a common origin;

determining a symbol rate for the signals in each channel;

descrambling encoded symbols in each channel with codes unique to the channel to enable the data to be extracted therefrom; and

multiplexing the descrambled data from the plurality of channels at a multiplexing rate derived from the symbol rates determined for the channels.

32. A method of receiving data as claimed in claim 31, further comprising decoding the multiplexed encoded data into frames of data.

33. A method of receiving data as claimed in claim 31 or 32, further comprising despreading the scrambled encoded symbols using a pseudo noise code.

34. A method of receiving data as claimed in claim 33, further comprising demodulating the despread symbols by way of variable decoding.

35. A method of receiving data as claimed in claim 34, wherein the despread symbols are demodulated with a respective walsh code.

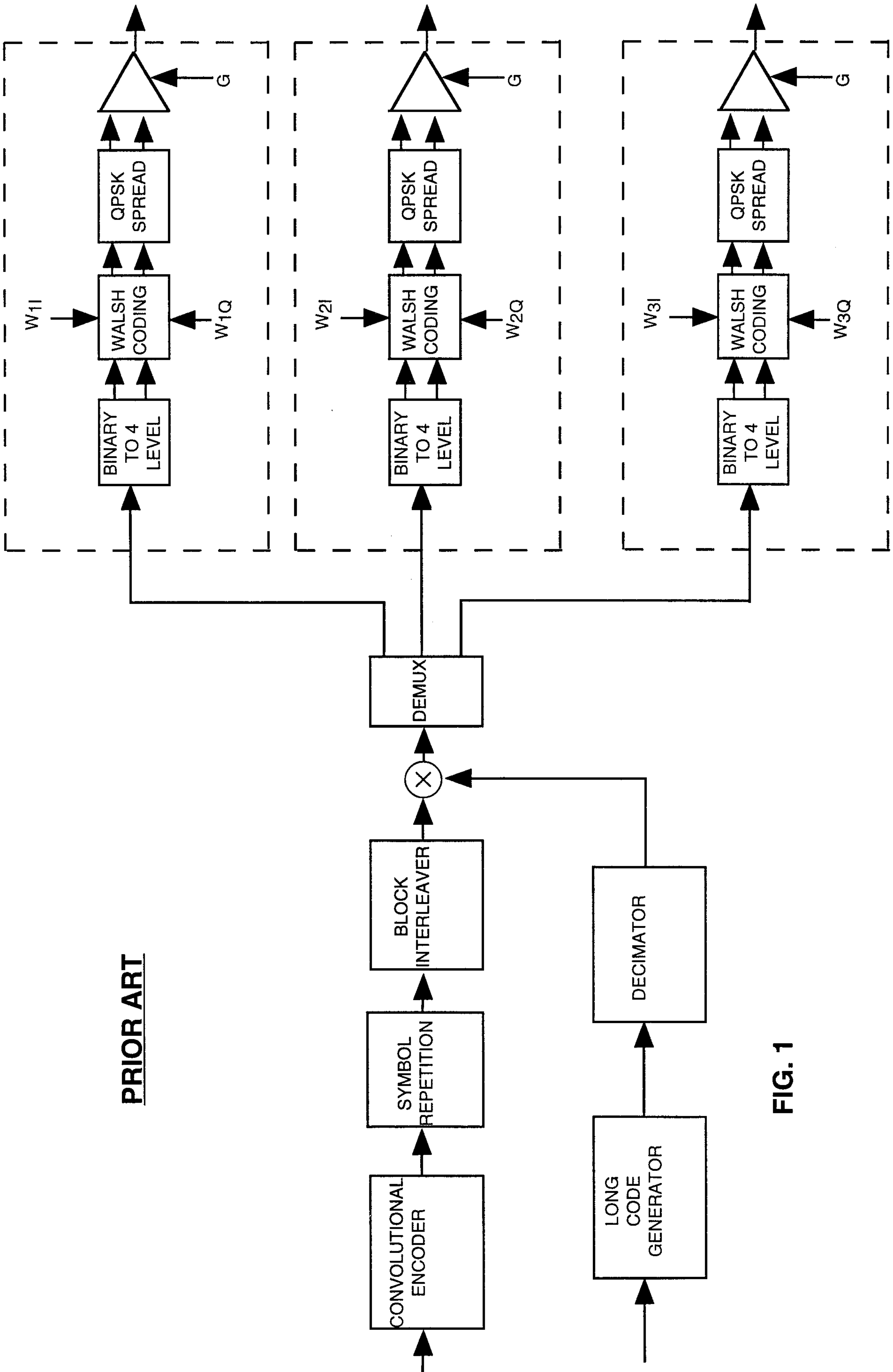
36. A method of receiving data as claimed in claim 35, further comprising descrambling the despread symbols in each channel using a respective long code for the channel.

37. A method of receiving data as claimed in claim 36, further comprising reordering the repeated symbols depending on a determined interleaving format.

38. A method of receiving data as claimed in claim 37, further comprising combining symbols in a channel before demultiplexing the same at a rate corresponding to the rate determined for the channel.

39. A method of receiving data as claimed in any of claims 31 to 38, further comprising:

demodulating the signals in different channels at different times.



PRIOR ART

FIG. 1

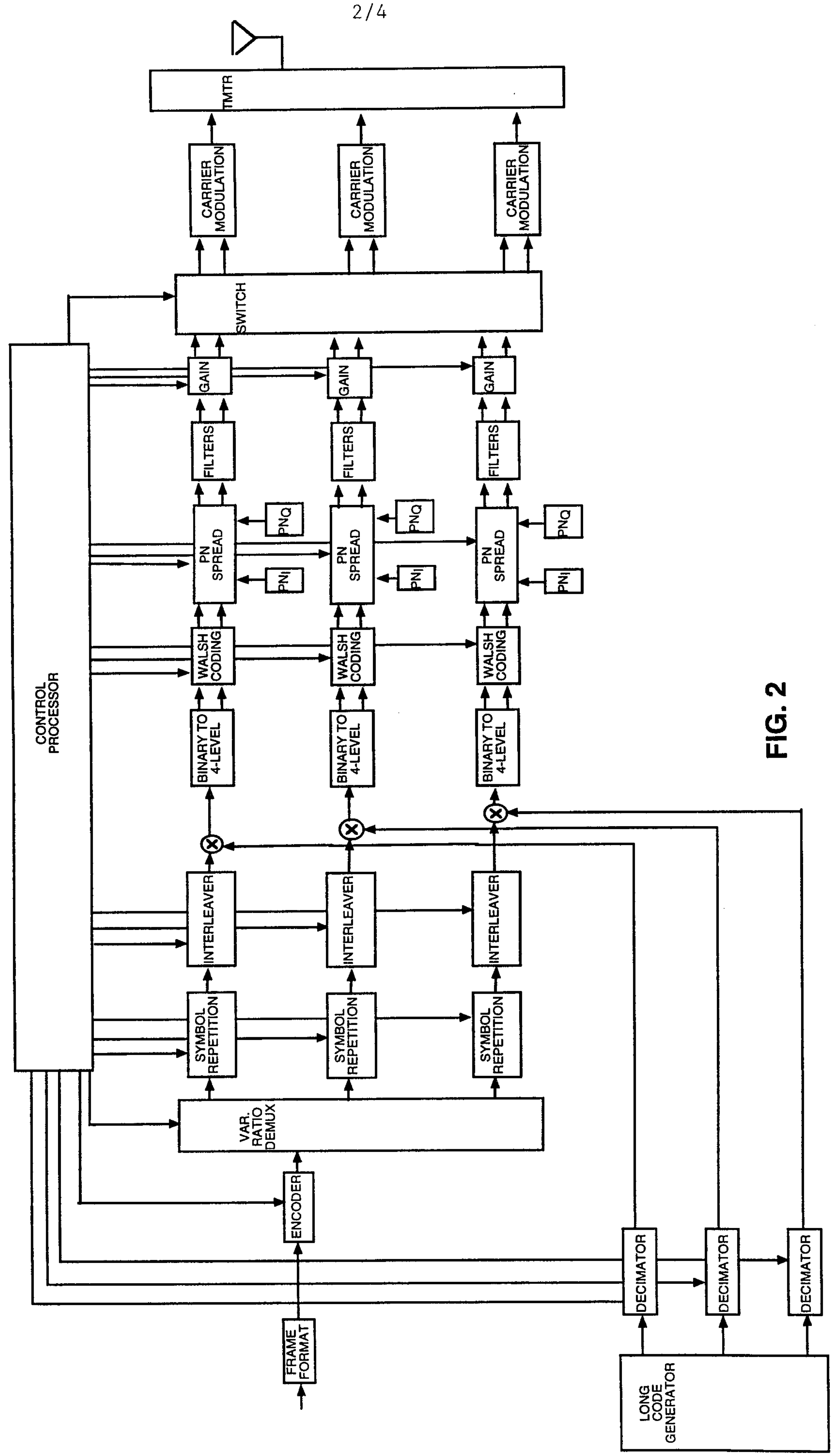

TMTR
CARRIER MODULATION
CARRIER MODULATION
CARRIER MODULATION
SWITCH
GAIN
GAIN
GAIN
FILTERS
FILTERS
FILTERS
PN SPREAD
PN SPREAD
PN SPREAD
PNQ
PNQ
PNQ
PNI
PNI
PNI
WALSH CODING
WALSH CODING
WALSH CODING
BINARY TO 4-LEVEL
BINARY TO 4-LEVEL
BINARY TO 4-LEVEL
INTERLEAVER
INTERLEAVER
INTERLEAVER
SYMBOL REPETITION
SYMBOL REPETITION
SYMBOL REPETITION
CONTROL PROCESSOR
VAR. RATIO DEMUX
ENCODER
FRAME FORMAT
DECIMATOR
DECIMATOR
DECIMATOR
LONG CODE GENERATOR

FIG. 2

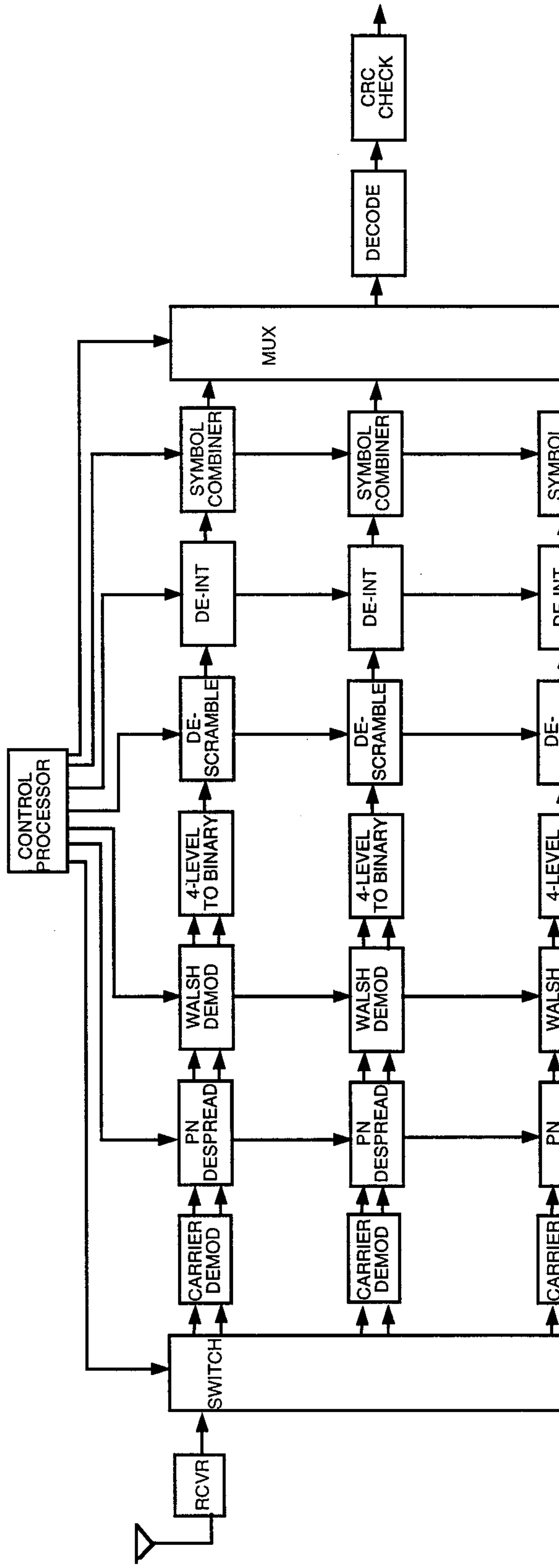

RCVR
SWITCH
CONTROL PROCESSOR
CARRIER DEMOD
PN DESPREAD
WALSH DEMOD
4-LEVEL TO BINARY
DE-SCRAMBLE
DE-INT
SYMBOL COMBINER
CARRIER DEMOD
PN DESPREAD
WALSH DEMOD
4-LEVEL TO BINARY
DE-SCRAMBLE
DE-INT
SYMBOL COMBINER
CARRIER DEMOD
PN DESPREAD
WALSH DEMOD
4-LEVEL TO BINARY
DE-SCRAMBLE
DE-INT
SYMBOL COMBINER
MUX
DECODE
CRC CHECK

FIG. 3

Walsh Chip within a Walsh Function

Walsh Function Index	0	1	2	3	4	5	6	7	8	9	10	11	12	13	14	15	16	17	18	19	20	21	22	23	24	25	26	27	28	29	30	31	32	33	34	35	36	37	38	39	40	41	42	43	44	45	46	47	48	49	50	51	52	53	54	55	56	57	58	59	60	61	62	63
0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0	0
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59	0	1	1	0	0	1	1	0	1	0	0	1	1	0	0	1	1	0	0	1	1	0	0	1	0	1	1	0	0	1	1	0	1	0	0	1	1	0	0	1	0	1	1	0	0	1	1	0	0	1	1	0	0	1	1	0	1	0	0	1	1	0	0	1
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61	0	1	0	1	1	0	1	0	1	0	1	0	0	1	0	1	1	0	1	0	0	1	0	1	0	1	0	1	1	0	1	0	1	0	1	0	0	1	0	1	0	1	0	1	1	0	1	0	0	1	0	1	1	0	1	0	1	0	1	0	0	1	0	1
62	0	0	1	1	1	1	0	0	1	1	0	0	0	0	1	1	1	1	0	0	0	0	1	1	0	0	1	1	1	1	0	0	1	1	0	0	0	0	1	1	0	0	1	1	1	1	0	0	0	0	1	1	1	1	0	0	1	1	0	0	0	0	1	1
63	0	1	1	0	1	0	0	1	1	0	0	1	0	1	1	0	1	0	0	1	0	1	1	0	0	1	1	0	1	0	0	1	1	0	0	1	0	1	1	0	0	1	1	0	1	0	0	1	0	1	1	0	1	0	0	1	1	0	0	1	0	1	1	0

FIG. 4